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## Contents



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# British Telecommunications Engineering



## EDITORIAL

On 6 August 1984, as a prelude to privatisation, the responsibilities, rights and liabilities of British Telecommunications were transferred to a public limited company—British Telecommunications plc—wholly owned by the government. Last year, British Telecom (the trading name of the company) had an annual turnover of some £6900M, with 23 000M calls being made on 29·3M telephones—approximately 4000M on the trunk network.

At the moment, this trunk network is largely analogue working, but rapid modernisation to an all-digital network by the end of this decade is planned. During this transition period, however, interworking between analogue and digital plant will be necessary, and a range of hybrid equipment, known as *transition equipment*, has been developed to provide these interconnections. A series of articles in this issue of the *Journal* introduces this equipment and describes the operation of transmultiplexers, codecs and data-in-voice modems and their application in the UK network.

The review of the performance standards of digital transmission systems in the July 1984 issue of the *Journal* is continued here with discussion on their specific application to satellite, metallic-pair and terrestrial-radio systems.

System X, another vital component in the modernisation of the UK network, since its introduction into service in 1980, has been evolving as technology has advanced. The next issue of the *Journal*, in January 1985, will be devoted to System X, and will reflect the evolution of the system and its role in the network of the future.

# Transition Equipment—An Overview

D. J. KINGDOM†

*Problems of interconnection between analogue and digital plant inevitably occur during the transition of a telecommunications network from analogue to digital methods of working. This article broadly reviews these problems together with some equipments specially developed to cope with them. This review serves as an introduction to three articles by other authors who describe some of these equipments in detail.*

## INTRODUCTION

The period of transition from analogue to digital methods of working in a telecommunications network is one in which the network will inevitably contain a complex and changing mixture of analogue and digital plant. Factors affecting this include the relative rates of change of switching and transmission systems, the types of equipment used on international connections, and the types of leased and customer-owned plant that need to be interconnected. A stage will be reached when, although quantities of working analogue plant will still exist within the network, all new installations and links will be of digital plant.

The successful application of digital techniques in telecommunications networks first took place on short-distance transmission systems using copper pairs to provide junction network type services. Their application to very long distance transmission, whether over copper or radio media, has taken place rather more slowly and, perhaps, will only reach its full potential on optical-fibre transmission media. Switching systems have adopted digital techniques at a rate, worldwide, somewhere between the two. For large countries, this means that digital switchers will have to be interfaced with analogue long-distance transmission systems for a long time to come. In the UK, however, very long distance transmission systems are found only on international links.

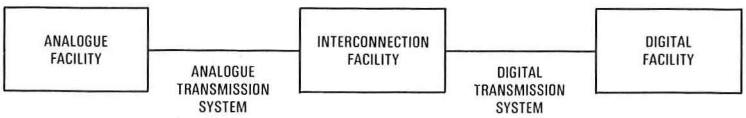
In coping with the transition problems in its national network, British Telecom (BT) has adopted an overlay strategy, which minimises the number of interfaces between analogue and digital plant, and a policy of rapid digitalisation, which minimises the transition time. Even so, in common with many other telecommunications administrations, the need to interconnect analogue and digital plant does not entirely disappear. The interconnections likely to be required by a telecommunications administration are broadly modelled in Fig. 1. Equipments that can provide such interconnections have been developed, and standards for them agreed within CCITT*. These equipments, which have become known collectively as *transition* equipment, comprise transmultiplexers, frequency-division multiplex (FDM) codecs (coders/decoders), data-in-voice modems and data-over-voice modems; the following articles[1, 2, 3] by Andrew Dick, Mike Andrews and Neil Harrison give detailed descriptions of some examples of these. Special equipment to provide the interconnection facilities shown in Fig. 1 is not always essential. In many cases, the job could be done by suitable back-to-back connections of standard FDM translating equipment and digital muldex (multiplex/demultiplex) equipment. The justification for special equipment is that it can do the job better and cheaper. Each type of equipment has a field of application to which it is most suited. This is discussed in the following sections.

## TRANSMULTIPLEXERS

This type of transition equipment transforms the signals from a number of channels assembled in one or more standardised FDM assemblies into the same number of channel signals assembled in one or more standardised time-division multiplex (TDM) assemblies, and carries out the reverse process. CCITT Recommendations cover two types

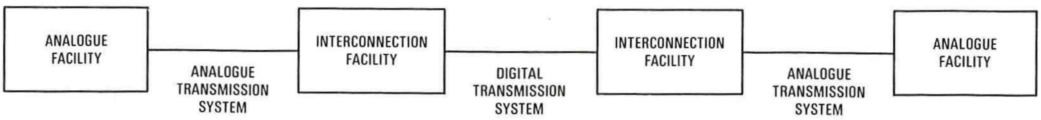
† Trunk Services, British Telecom National Networks

* CCITT—International Telegraph and Telephone Consultative Committee




*Note*: Either the analogue or the digital transmission system could reduce to station cabling

(*a*) Use of transmultiplexers as the interconnection facility




*Note*: Either or both of the analogue transmission systems could reduce to station cabling

(*b*) Use of FDM codecs as the interconnection facility


*Note*: Either or both of the digital transmission systems could reduce to station cabling

(*c*) Use of data-in-voice or data-over-voice modems as the interconnection facility

FIG. 1—Examples of interconnections between analogue and digital systems

of transmultiplexer. The first can provide interconnection between 24 channels assembled in two basic groups and 24 channels assembled in one 1544 kbit/s primary pulse-code modulation (PCM) group. The second can provide interconnection between 60 channels assembled in a basic supergroup and 60 channels assembled in two 2048 kbit/s primary PCM groups. These two bit streams can be mutually plesiochronous. Transmultiplexers can, therefore, provide the interconnection facilities shown in Fig. 1(*a*) where typically the analogue and digital facilities would be switchers in a public or private telephone network. Access to individual channels can be made at the analogue-facilities end by means of standard FDM translating equipment, and at the digital-facilities end, if necessary, by means of standard digital muldex equipment.

Early demands for transmultiplexers arose in the USA where the first versions of the equipment consisted essentially of channel translating equipment connected at audio frequencies to PCM muldex equipment. Progress in integrated-circuit technology has made possible more elegant solutions in which real-time digital signal processing is used to transform the channel signals without the need for an audio stage. Compared to the early equipments, this technology has improved transmission performance, stability with time, size, power consumption and cost. Andrew Dick's article describes a 60-channel digital signal processing transmultiplexer and outlines the concepts on which this type of equipment is based. They are not simple; more that $10^6$ multiplications per second have to be carried out for each channel in addition to other arithmetic and storage operations. The largest transmultiplexers commercially available at present seem to be equipment for processing 60 channels.

Although the audio stage is eliminated in the digital signal processing type of transmultiplexer, processing is still carried out on each 4 kHz channel signal as defined for the FDM and TDM assemblies. This means that 3 kHz channelling and services such as sound programme circuits, which require bandwidths greater that 4 kHz, cannot be carried by these transmultiplexers. Signalling presents further problems; a lack of correspondence between digital and analogue signalling systems means that general solutions are not possible. A means of coping with CCITT Signalling System R2 has been agreed within CCITT and should be published in 1985. Generally, transmultiplexers provide good interconnection facilities for voice services carried on 4 kHz FDM channels and 64 kbit/s PCM-encoded TDM channels.

To date there has been no identified need for transmultiplexers in BT's inland network—it will be remarkable if no such need ever arises. Andrew Dick explains a particular application for international working that may be regarded as a complex version of the example given in Fig. 1(*a*).

## FDM CODECS

A pair of FDM codecs make it possible for an FDM channel assembly to be carried over a digital transmission system and so provide the interconnection facilities shown in Fig. 1(*b*). The analogue facilities shown in the figure may be envisaged as switchers in a public or private network, as sound programme equipment or as customer-owned analogue data communications equipment. At the send terminal of the digital link, the analogue multichannel signal is encoded as a whole, without the signal being sectionalised in any way, into a bit stream suitable for transmission over that digital link. At the receive terminal, the digital signal is decoded to recover the analogue multichannel signal. A CCITT Recommendation for FDM codecs is due for publication in 1985. It is expected to recommend that the FDM assemblies to be encoded should be standardised assemblies, for example, basic supergroup, mastergroup, 15-supergroup assemblies etc, and will detail the overall analogue-to-analogue performance required. There will be no recommendation made as to encoding strategy or frame structure: decisions on these will be left to individual designers. In so far as the digital signal is concerned, the CCITT will recommend only that it be suitable for transmission over a digital transmission system operating at a standardised hierarchical bit rate. This makes it necessary for administrations to ensure that compatible designs of FDM codec are used to form a pair on each link.

Mike Andrews' article describes two variants of FDM codec: one encodes a basic supergroup for transmission over an 8·448 Mbit/s link, and the other encodes a basic 15-supergroup assembly† into a 68·736 Mbit/s bit stream. In the latter case, two plesiochronous 68·736 Mbit/s streams can be combined for transmission over a 140 Mbit/s link. Although 68·736 Mbit/s is not a CCITT recommended hierarchical bit rate, it is standard practice for BT to provide an access port at this bit rate on its 34–140 Mbit/s muldex equipment.

FDM links provided by means of FDM codecs have noise characteristics different from those of links provided by means of all-analogue plant. For a pair of FDM codecs, the proposed CCITT Recommendation is for a noise allowance of 800 pW0p* maximum per 4 kHz channel with the expectation that only FDM codecs for the larger channel assemblies will take up all of this allowance. In practice, 15-supergroup assembly codecs have achieved a noise performance of about 500 pW0p, while supergroup codecs have achieved 150 pW0p. Errors and jitter on the digital link add to this noise, but under normal conditions the increase should be negligible. An FDM link provided by means of FDM codecs, therefore, has a fixed noise performance, determined by the terminal FDM codecs, and is independent of the length of the associated digital transmission system. In contrast to this, for analogue national links, the appropriate CCITT noise allowance is 3 pW0p/km for line links and 60 pW0p for translating equipment. A 15-supergroup assembly link provided over a 12 MHz line system and a pair of 15-supergroup assembly translating equipments may be regarded as interchangeable with a similar link provided by means of a 15-supergroup assembly codec at each end of a digital line system. The noise break-even point of 500 pW0p occurs with a 12 MHz line system length of about 150 km. For a supergroup link, a comparison can be made between a link provided by means of a supergroup codec at each end of a digital line system and a similar link provided by means of a 12 MHz line system, a pair of 15-supergroup assembly translating equipments and a pair of supergroup translating equipments. The noise break-even point of 150 pW0p occurs with a 12 MHz line system length of 10 km.

The ratio between the number of 4 kHz channels that can be carried by the bit stream derived from an FDM codec and the number of channels that can be carried by the same bit rate derived from all-digital muldex equipment can be regarded as the utilisation efficiency of that FDM codec. It is always less than unity. In the case of the hypergroup codec described by Mike Andrews, the ratio of channels is 900/960. A less favourable ratio of 60/120 is achieved by the supergroup codec.

It is possible to use a pair of transmultiplexers to provide the interconnection facilities shown in Fig. 1(*b*) instead of FDM codecs, but on a number of counts this may be unattractive. FDM codecs use well-known techniques and, compared with transmultiplexers, are simple and cheap. One FDM codec at each end of a link is capable of carrying 900 channels. If 60-channel transmultiplexers were used, 15 equipments would be required at each end of a link to process 900 channels in addition to some higher-order digital muldex equipment. There is also the point that digital signal

† In the UK, a 15-supergroup assembly is normally referred to as a hypergroup. The two terms are used interchangeably in this article

* Psophometrically weighted noise power in picowatts referred to a point of zero relative level

processing transmultiplexers can carry only 4 kHz channels whereas FDM codecs have no such restriction because they encode the analogue multichannel signal as a whole. These codecs can carry 3 kHz channelling, sound programme signals and wideband analogue data, and present no problems with signalling. Against this has to be set the utilisation efficiency of the FDM codec.

For the hypergroup codec, this is likely to be a significant factor only on the most expensive long-distance digital link. For 60 channels, although an FDM codec is likely to be cheaper than a transmultiplexer, the utilisation efficiency of 60/120 for the FDM codec will have a greater effect on any financial assessment of the alternatives. If transmultiplexers are used to provide this type of interconnection, signalling, if any, needs consideration. In-band signalling and common-channel signalling, provided they are within the band 300–3400 Hz, should present no problems. R2 signalling could be used with conversion at each transmultiplexer; any other forms of out-band signalling require specific solutions to be devised. Similarly, reference pilots for any of the analogue assemblies cannot be carried through the digital link.

Modest use of both supergroup and 15-supergroup assembly codecs are planned by BT, with the latter equipment being the more important.

## DATA-IN-VOICE AND DATA-OVER-VOICE MODEMS

These somewhat curiously named equipments enable a digital bit stream to be carried over an analogue transmission system and can, therefore, provide the interconnection facilities shown in Fig. 1(*c*). The digital facilities shown in Fig. 1(*c*) are typically digital data equipment in a public or private network. In the case of systems using data-in-voice modems, the digital signal is carried within the frequency band normally occupied by a standardised FDM assembly; for example, the equipment described in Neil Harrison's article allows a 2·048 Mbit/s signal to be carried in a frequency band normally occupied by two contiguous supergroups in a hypergroup. In systems using data-over-voice modems, the digital signal is carried by the analogue transmission system in a frequency band above that occupied by the standardised FDM assemblies. This is possible because most analogue transmission systems have some useable bandwidth above the highest frequency used for the FDM assemblies and above the frequency of any system pilot. Although the noise performance in this band is likely to be outside the limits required for analogue transmission, it may still be good enough to allow acceptable performance with digital-type signals. Some transmission systems already use this band for supervisory signals, and it may not, therefore, be available without some re-engineering.

Essential requirements for both data-in-voice and data-over-voice systems are modulation and frequency-translation processes to convert the digital signal into a form suitable for transmission over an analogue link, and a coherent demodulation process to provide correct reconstitution of the digital signal waveform. No particular processes are recommended by CCITT. Techniques which have been used include phase-shift keying and vestigial-sideband amplitude modulation. Administrations must ensure that compatible designs of equipment are used at each end of the analogue line system.

At first sight, the proposition to carry digital signals over an analogue multichannel transmission system may not seem very attractive. Economic dimensioning of the load-carrying capacity of amplifiers used in FDM systems necessarily allows for a small but finite probability of overload during which the peaks of the multichannel signal are clipped[4]. Also, there has been little apparent need to rigorously engineer out all sources of impulsive-type noise found in analogue systems. In both cases, the impairments caused to analogue voice signals are small, but this is not necessarily the case with digital signals, particularly those carrying digital data services. However, in practice, overload of FDM systems occurs far less frequently than is assumed in design objectives, and most sources of impulsive noise can be tracked down and eliminated. With care taken in loading the analogue system with the signal from the data-in-voice or data-over-voice modems, a useful digital service can be provided. The systems used need to be well maintained to prevent level misalignments and to prevent noisy groups or supergroups generating interference. Perhaps the greatest care in the application of these modems needs to be taken on international links, where the use of 3 kHz channelling, high levels of data-type traffic, compandors and circuit multiplication equipment all lead to high loading levels on FDM systems.

Utilisation efficiency for data-in-voice systems may be defined in a similar manner to that used for FDM codecs. It is always rather poor and, in the case of a 2·048 Mbit/s signal carried over two supergroups, is 30/120. A pair of transmultiplexers may then be considered as an alternative way of providing this type of interconnection; but again there would be problems caused by the channel processing of the transmultiplexers. Data-in-voice modems provide true bit transparency over the analogue path. The digital bit stream can carry sub-64 kbit/s encoded speech, high bit rate data or digitally-encoded sound programme signals. None of these can be carried by digital signal processing transmultiplexers.

BT has not used data-over-voice systems because of the difficulty of implementation on their existing analogue line systems. Data-in-voice systems have been used, in a very small way as yet, to provide digital services in advance of digital transmission facilities being available. Despite their poor utilisation efficiency, they can be very useful in this application, although their use by BT is unlikely ever to become widespread.

## CONCLUSIONS

Some of the problems of interconnection between analogue and digital plant that arise during the digitalisation of an existing analogue telecommunications network can be solved by the correct application of transmultiplexers, FDM codecs, data-in-voice modems and data-over-voice modems. Each has an appropriate area of application. No doubt, in some cases, the choice may not be obvious and requires careful consideration of the limitations of each equipment as well as the facilities they offer. These specialised equipments have been evolved to meet certain analogue-to-digital interconnection requirements which will disappear when the network in which they are used becomes all digital; at this point these equipments should become extinct.

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[3] HARRISON, N. A 2048 kbit/s Data-In-Voice Modem. *ibid.*, Oct. 1984, **3**, p. 161.

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### Biography


Dennis Kingdom is a section head in the Trunk Engineering Division of BT National Networks. He joined what was then the Engineer-in-Chief's office from the Bournemouth Telephone Area in 1956. Since then, he has been involved in a number of developments in the transmission field, including transition equipments. He is currently head the of the UK delegation to CCITT Study Group XV, which studies transmission systems.

# Transmultiplexers

A. B. DICK, B.SC.†

UDC 621.38.037.37 : 621.38.037.33

*This article describes the network and technical aspects of a 60-channel transmultiplex equipment. The possible applications within the network and the problems involved in incorporating the equipment are discussed, and a brief description of the mathematical algorithm used to achieve the transmultiplexer function is given.*

## INTRODUCTION

The name *transmultiplexer* (TMUX) is a general term used to describe any equipment that directly converts a signal assembled by frequency-division multiplex (FDM) equipment to a signal that can feed into standard pulse-code modulation (PCM) multiplexing equipment. CCITT* describe this as follows: 'a TMUX transforms FDM signals such as a group or supergroup into TDM signals that have the same structure as those derived from PCM multiplex equipment and also carries out the reverse process'. To do this transformation, a TMUX must perform a channel-by-channel conversion. Consequently, it can carry only voice-frequency services; services such as 64 kbit/s data, group-band data and music-in-band cannot be carried through a TMUX.

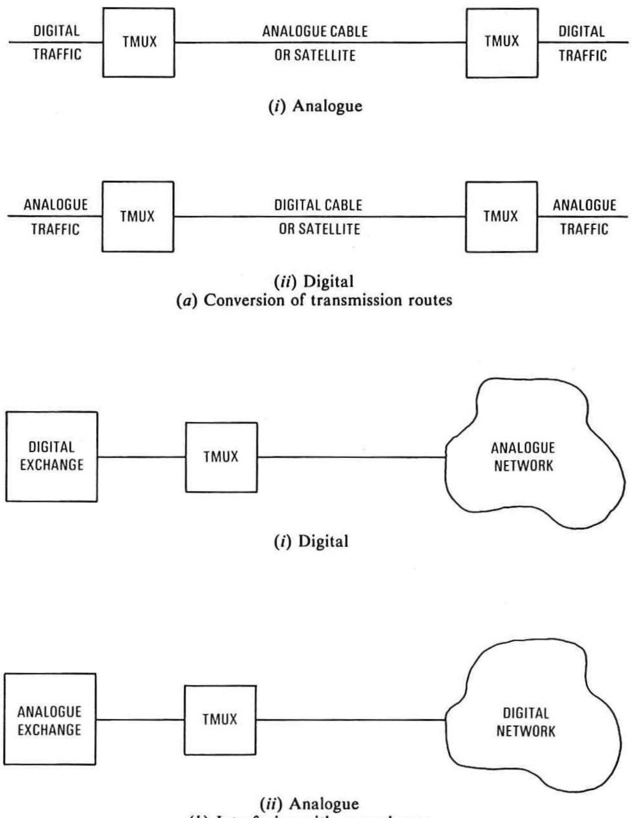
In practice, there are two types of TMUX: a 24-channel version, which is designed to suit 1·544 Mbit/s 24-channel PCM, and a 60-channel version, which is designed to suit 2·048 Mbit/s 30-channel PCM. The main users of 24-channel PCM are in North America and Japan and, hence, it is there that 24-channel TMUX is of interest. In Europe and Australasia, 30-channel PCM is standard, and so it is the 60-channel TMUX that is of interest: it is this version that is described in this article. The 60-channel TMUX converts an FDM supergroup into two 2·048 Mbit/s PCM streams.

The TMUX can act as an interface device between the existing analogue network and the new evolving digital network. British Telecom International (BTI) is implementing a policy to use these devices to provide this interface where possible, and has placed an initial order for equipment to be in service later this year.

## USE OF TRANSMULTIPLEXERS IN TELECOMMUNICATION NETWORKS

The TMUX is designed to be an interface device between an analogue network and a digital network and, therefore, can be used in a number of situations. The four main possibilities are illustrated in Fig. 1.

With the distinction between transmission and switching, the TMUX can be seen to do two separate functions. First, it is a means by which a transmission link can be used the other way round; that is, an existing analogue link that can carry digital telephony and datel services by the use of a TMUX at each end. Conversely, if a new digital link has been provided and if the carrying of analogue traffic is still required, then this service can be provided by using TMUX at each end of the link. The second function can be seen by considering the switching aspects. If a new digital switch is provided at a site, then it may be required to interface with existing analogue routes. By using the TMUX, the switch can be made to look like an analogue exchange. Similarly, an analogue switch can be made to look like a digital exchange by using a TMUX as the interface. The particular use to which a TMUX is put depends largely on the policy adopted by telecommunication administrations to change their network from analogue to digital. One of the important factors is whether digital exchanges are deployed before, after or at the same time as digital line plant.




(*i*) Analogue
(*ii*) Digital
(*a*) Conversion of transmission routes
(*i*) Digital
(*ii*) Analogue
(*b*) Interfacing with an exchange

FIG. 1—Uses of transmultiplexers

## USE OF TRANSMULTIPLEXERS BY BTI

In the international network, any one telecommunication administration cannot have control over the whole network; so BTI has had to adopt a more flexible policy with regard to interfacing analogue to digital transmission than is pursued by British Telecom nationally. Because of the particular type of services BTI provides, the switching centres are large and centralised. This type of exchange is usually provided most economically if digital switching is used. A large proportion of international traffic is carried on submarine cables, which are very expensive to lay, and which are expected to remain in service for 25 years. Until very recently, all these cables have been analogue working. Thus, BTI is left in the position of having to interwork digital switching and analogue line plant for some period of time.

† International Lines, British Telecom International
* CCITT—International Telegraph and Telephone Consultative Committee

FIG. 2—Transmultiplexer installation at Keybridge House

At present, BTI has an operational digital switch at Keybridge House, but, as yet, no digital international transmission is available. The first digital transmission system to another country is planned to be the European Communications Satellite (ECS), which has been launched and is, at present, undergoing acceptance testing. This has led to the use of the TMUX by BTI, both to provide an initial means of restoring service should ECS fail and, in the longer term, as a general means of interfacing the Keybridge switching unit to the existing analogue satellites and cables. Other countries are finding that economics can make TMUX an attractive proposition. It can now be cheaper to buy a digital exchange and completely interface it with analogue transmission systems by using TMUX than it would be to buy an analogue switching system.

Two main factors appear to have delayed the deployment of the TMUX. Firstly, telecommunication administrations have been slow to define their exact technical requirements for the TMUX. This has meant that manufacturers have been slow to produce a device that can be easily configured for a number of network situations; for example, providing a number of options for pilots, signalling systems, etc. Secondly, the technology used by most manufacturers to make the TMUX is very advanced, and administrations have been sceptical of both its performance and reliability as well as the implications of this new technology on network performance and maintenance procedures.

These arguments should no longer be valid and, in order to prove that the TMUX can be incorporated in the network, BTI undertook a trial of TMUX equipment. This trial has shown that the TMUX is very reliable and that the TMUX can be put into the network so that it is completely transparent to normal voice-band services. The trial not only tested the passing of speech and datel services through the transmission path, but also the operation of a signalling converter. This converter can transform an international out-of-band signalling system into an equivalent system which uses time-slot 16 (TS16) of that PCM frame. It is designed to be transparent, and during the trial it showed no tendency to lock up the routes. In all, much confidence has been gained in the use of the TMUX.

## TRANSMISSION REQUIREMENTS OF TMUX

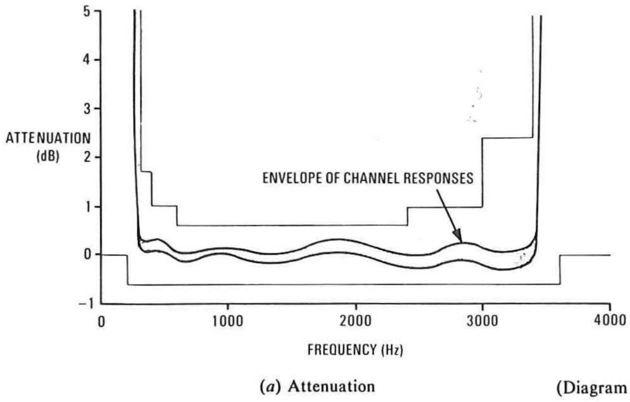
The transmission requirements of the TMUX have been specified by CCITT in Recommendation G792. This can be broken down into three main areas:

(*a*) analogue interfaces—supergroup,
(*b*) digital interfaces—2 Mbit/s, and
(*c*) transmission performance through TMUX.

The interface specifications are based on standard analogue and digital Recommendations[1, 2, 3]. Since the TMUX is inserted into the network at a point where channel filtering would not normally be carried out, the transmission performance does have to be specified closely.

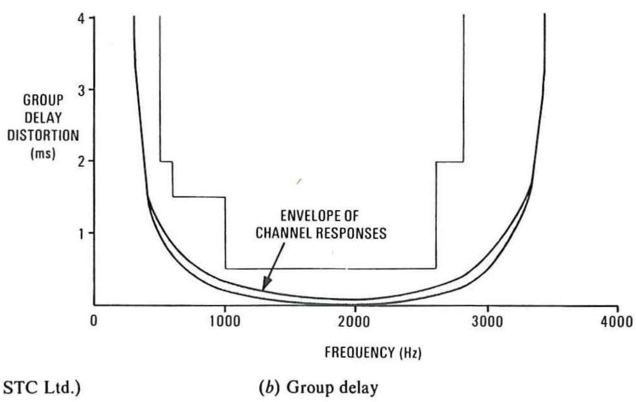
Most manufacturers have adopted a technology called *digital signal processing* (DSP) in order to implement a TMUX, and one of the advantages of this technique is that the channel is very accurate and very reliable. Fig. 3 shows the channel responses for all 60 channels on the equipment supplied for the trial. As can been seen from Fig. 3, they are all virtually identical. Because most of the performance features of systems that use DSP show the same type of imperfections as their analogue equivalents, parameters such as idle-channel noise, linearity, intermodulation, and even crosstalk have to be specified and carefully measured as with any analogue equipment.

The main source of noise within any digital equipment is from quantisation distortion. This arises because the signal is approximated to specific values. (This is equivalent to adding a noise source to the signal.) Thus, the performance of the TMUX with respect to idle-channel noise, quantisation distortion, linearity, intermodulation etc. depends on the design of the analogue interface and on the design of the




(*a*) Attenuation

(Diagram courtesy STC Ltd.)




(*b*) Group delay

FIG. 3—Frequency and group-delay distortion on all 60 channels of BTI trial transmultiplexer

## TABLE 1
## Limits Specified in Recommendation G792

| Parameter | Limit | Recommendation G792 |
|---|---|---|
| Idle channel noise | −65 dBm0p | Paragraph 11.1 |
| Total distortion (Note 1) | 32·5 dB | Paragraph 13 |
| Linearity | 0·5 dB | Paragraph 15 |
| Intelligible crosstalk | −62 dB | Paragraph 16.1 |
| Unintelligible crosstalk | −60 dB | Paragraph 16.2 |

*Note* 1: Includes quantisation distortion

DSP section. DSP relies on the arithmetic manipulation of numbers and leads to another source of noise, which arises because arithmetic, when worked with only a limited number of digits, produces a rounding error. When many calculations are carried out, this error can grow, and in DSP gives rise to noise. This can be very substantially reduced by performing the calculations with a greater precision than that of the original or final numbers. A list of some of the limits specified in CCITT Recommendation G792[4] is given in Table 1.

### SIGNALLING

The number of places where a TMUX can be effectively used in the network can be greatly increased if the TMUX has the facility to convert signalling information from an analogue signalling system to its digital equivalent. European countries have agreed that, initially, digital service using a variation of the current analogue R2 signalling system[5] should be provided. The analogue version of R2 signalling uses an out-of-band tone (3825 Hz) to carry the line signalling while the register signalling uses in-band multi-frequency (MF) tones. The digital version of R2 signalling uses the same in-band register signalling as analogue R2, but uses TS16 of the PCM frame to carry the line signalling. This is done by using the standard multiframe structure specified in CCITT Recommendation G732. The line signalling codes for each are shown in Table 2. In order to use the TMUX effectively in the international network, the facility to convert the digital line signalling in TS16 to analogue out-of-band tones is required.

The register signalling, being in-band, passes through the transmission path of the TMUX and so needs no conversion. On the other hand, the 3825 Hz tones have to be extracted from the FDM multiplex and the information then converted into the TS16 bits. The extraction of the tones from the FDM multiplex can be done by using the same DSP technology as is used for the basic multiplexing and demultiplexing operations. The logic needed to do the conversion requires very fast microprocessor technology in order to cope with all 60 channels. The sequence that each channel must follow is specified by the CCITT[6].

## TABLE 2
## R2 Line Signalling Codes

| Signal | Digital ($a$, $b$ bits TS16) | | | | Analogue (3285 Hz Tone) | |
|---|---|---|---|---|---|---|
| | $a_f$ | $b_f$ | $a_b$ | $b_b$ | $t_f$ | $t_b$ |
| *Idle* | 1 | 0 | 1 | 0 | 1 | 1 |
| *Seizure* | 0 | 0 | 1 | 0 | 0 | 1 |
| *Seize acknowledge* | 0 | 0 | 1 | 1 | 0 | 1 |
| *Answer* | 0 | 0 | 0 | 1 | 0 | 0 |
| *Clear back* | 0 | 0 | 1 | 1 | 0 | 1 |
| *Clear forward* | 1 | 0 | 0 | 1 | 1 | 0 |
| | | | or 1 | 1 | or 1 | 1 |
| *Blocking* | 1 | 0 | 1 | 1 | 1 | 0 |

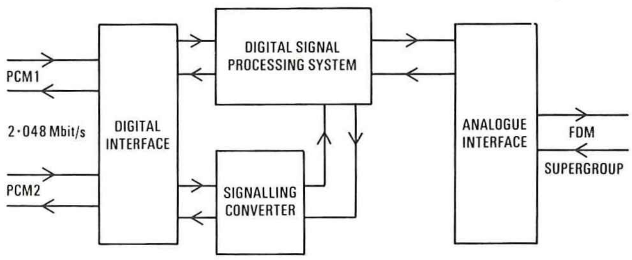
f: forward b: backward

Other possibilities for converting signalling information include the extraction of in-band 2280 Hz tones as well as the extraction of the whole TS16 when it is being used to carry common-channel signalling. This is extracted as a 64 kbit/s digital stream, and can either be diversely routed on the KiloStream network or transmitted over the analogue network by using a group band modem.

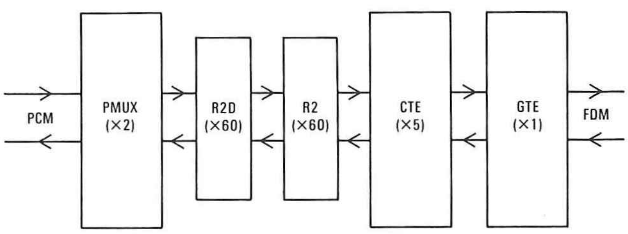
### THE DESIGN CONCEPT—AN ALGORITHM

The basic idea employed by most manufacturers avoids any conventional analogue audio paths in the equipment and dispenses with the traditional analogue multiplexing techniques (see Fig. 4). Instead, the PCM signals are kept in digital form and all the filtering and multiplexing is done on these digital signals. The final digital signal is then converted directly into the FDM multiplex. The digital signal is a series of numbers which pass through the equipment at a certain rate. The rate is known as the *sampling rate* and is one of the fundamental properties of the signal.

Just as it is possible to process analogue signals—amplify, attenuate, filter, modulate etc.—so it is possible to do the same with this train of numbers or digital signal. This is called *digital signal processing.* For example, an analogue filter can be designed by using the conventional frequency response techniques which produce a circuit containing resistors, inductors, and capacitors. It is equally possible to specify the frequency response required of a digital filter which will produce a design containing adders, multipliers, and registers which will delay the digital signal (a train of numbers) for a certain number of sample periods. A simple third-order digital filter could be implemented by using the system shown in Fig. 5. The computational process that specifies the digital filter is called an algorithm. The techniques used to design DSP systems are very similar to those used to design analogue systems. As with analogue systems, a complex transform method is used. In order to design an analogue filter, a function of the complex variable $s$ which describes the characteristics of the filter is derived. This function is arrived at from the frequency or phase response




*Note*: This is equivalent to the following if conventional transmission equipment were used




PMUX : 30-channel primary multiplex
R2D : Digital R2 relay-set
R2 : Analogue R2 relay-set
CTE : Channel translating equipment (12 channel)
GTE : Group translating equipment

FIG. 4—Block diagram of a transmultiplexer

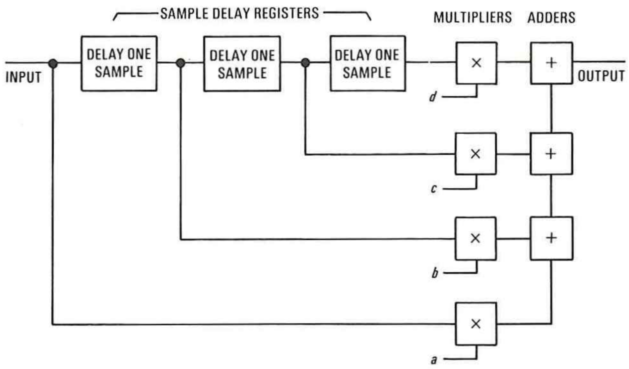



FIG. 5—Simple third-order digital filter

required of the the filter, and from it can be found the exact circuit required in order to implement the filter. To design a digital filter, a function of the complex variable $z$ is derived in exactly the same way. It is first of all useful to note that the two variables $s$ and $z$ are closely related.

$$z = e^{sT}$$

where $T$ = sampling period

Secondly, since the signals are in the form of a train of numbers, the signal can be described by a number series rather than by a mathematical function. This number series, like a function, is normally given a letter to describe it, say $g$, and $g_k$ is used to describe the $k$th number in the series. So for an input signal $g$ and an output signal $r$, it is possible to define the transforms of these signals.

$$G(z) = \sum_{k=-\infty}^{\infty} g_k z^{-k}$$

$$R(z) = \sum_{k=-\infty}^{\infty} r_k z^{-k}$$

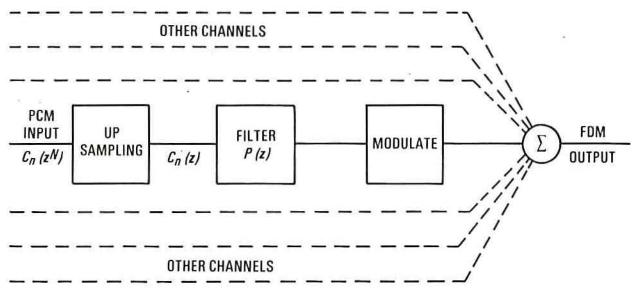
The filter can now be described as a function of $z$ say $H(z)$ which relates $G(z)$ to $R(z)$.

$$R(z) = H(z)G(z).$$

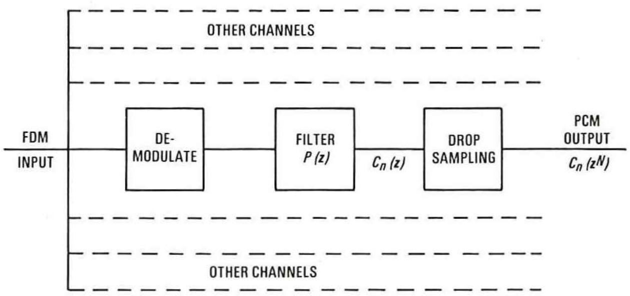
The main advantage of using $z$ for the transform is that $z^{-k}$ means 'delay the signal $k$ samples'. The filter shown in Fig. 5 can therefore be described by a $z$ transform as follows:

$$\begin{aligned} R(z) &= aG(z) + bz^{-1}G(z) + cz^{-2}G(z) + dz^{-3}G(z) \\ &= (a + bz^{-1} + cz^{-2} + dz^{-3})G(z) \\ H(z) &= a + bz^{-1} + cz^{-2} + dz^{-3} \end{aligned}$$

The filtering and modulation required by the TMUX can be defined by these $z$ transforms. The PCM channels on one side of the TMUX are sampled at 8 kHz and so can be directly described by a $z$ transform where the sampling period is 125 $\mu$s. The FDM side must be sampled at a much higher rate of $8 \times N$ kHz, where $N$ is some integer number which is greater than the number of channels in the FDM multiplex. This is necessary to ensure that no aliasing occurs on the FDM side. By taking different values for $N$, various advantages can be gained, and almost as many choices are available as there are manufacturers. The manufacturer who supplied the equipment for the BTI trial takes $N$ to be 14, since this means that no frequency conversion stage is necessary at group level. Other manufacturers have taken $N$ to be 64 because this leads to a simpler algorithm; others have taken $N$ to be 72, since this requires no frequency conversion stage on the supergroup. The manufacturer of the trial equipment is planning a second generation TMUX which will take $N$ to be 70. This will lead to a simplification of hardware, but will still be compatible with the original design. It appears that one solution is no better than any other, and that the performance of the TMUX is not really affected by this choice, but by other more peripheral aspects of the design. Within the TMUX, there are two sampling frequencies, one for the FDM and one for the PCM. It is convenient to define one by $z$ and relate the other one to it. The FDM is taken as the base sampling rate.




(*a*) PCM-to-FDM multiplex




(*b*) FDM-to-PCM demultiplex

FIG. 6—Block diagram of the operation of a transmultiplexer

$$\text{for FDM} \qquad z = e^{sT}$$

$$\text{for PCM} \qquad z^N = e^{sNT}$$

where $T$ = FDM sampling period, and
$NT$ = PCM sampling period.

The TMUX can now be thought of as shown in the block diagram in Fig. 6.

Now that the basic design has been established, it is possible to look for simplifications to reduce the amount of computation required. It is possible to reduce substantially the number of calculations required by manipulating the mathematics of the filter $P(z)$ by decomposing it into the sum of $N$ filters each working at the PCM sampling rate.

$$P(z) = \sum_{k=0}^{N-1} P_k(z^N)z^{-k}$$

The end result is two formulae which describe the TMUX, one for the PCM-to-FDM multiplex direction and the other for the FDM-to-PCM demultiplex direction.

PCM to FDM multiplex

$$M_k(z^N) = \sum_{n=0}^{N-1} C_n(z^N)P_k(z^N)e^{j2\pi nk/N}$$

where $M_k(z^N)$ = $k$th sample of FDM
$C_n(z^N)$ = $n$th PCM channel
$P_k(z^N)$ = filter
$e^{j2\pi nk/N}$ = modulate

FDM to PCM demultiplex

$$C_n(z^N) = \sum_{k=0}^{N-1} M_k(z^N)e^{-j2\pi nk/N}P_{-k}(-z^N)$$

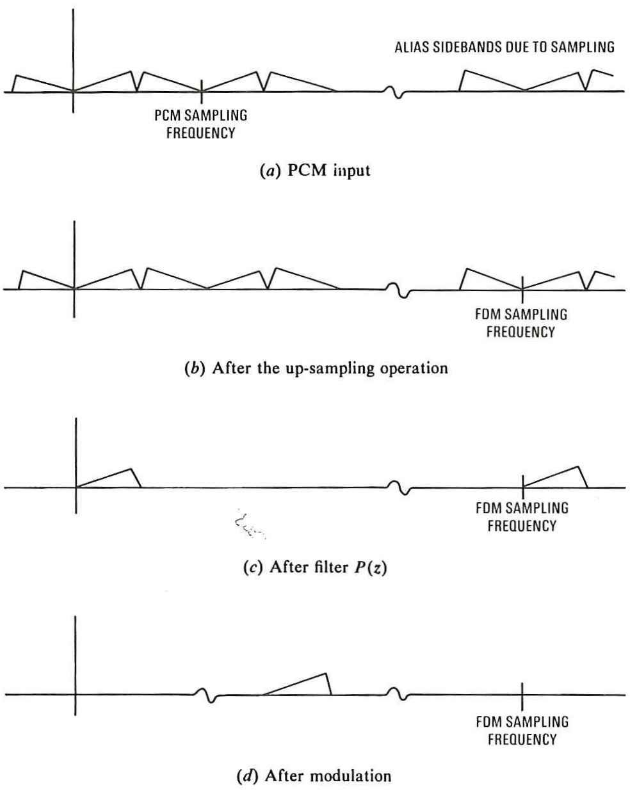



FIG. 7—Frequency spectra during TMUX operation in PCM-to-FDM direction

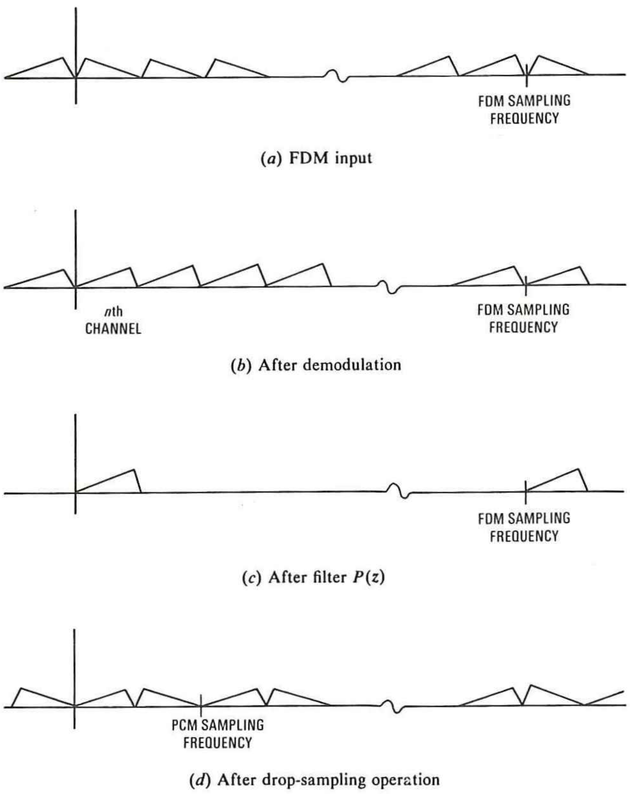



FIG. 8—Frequency spectra during TMUX operation in FDM-to-PCM direction

where $C_n(z^N)$ = $n$th PCM channel
$M_k(z^N)$ = $k$th sample of FDM
$e^{-j2\pi nk/N}$ = demodulate
$P_{-k}(-z^N)$ = filter

The spectrum of the signal as it goes through these various stages is shown in Figs. 7 and 8.

This technique of using digital filtering to carry out multiplexing and demultiplexing was first described by Darlington[7] in 1970 as a mathematical idea long before digital hardware was available to implement the technique. In the interim period, many manufacturers as well as telecommunication administrations and academic institutions have developed variations on the idea which optimise the algorithm for various parameters[8]. Since these algorithms are purely digital, their performance can be simulated on digital computers where the calculations that are carried out are exactly the same but carried out at a much slower rate than in a real TMUX. These variations have all been developed and tested on digital computers to find the basic filter characteristics. These computer models also predict other parameters like quantisation distortion, linearity and crosstalk.

The concepts of DSP are the basis now of all TMUX designs as well as the basis for other types of equipment; for example, echo cancellers; adaptive differential PCM (ADPCM), where speech is carried at 32 kbit/s or even 16 kbit/s; 2 Mbit/s videoconferencing; and circuit multiplication equipment (CME). This area of technology is only just beginning to be explored, but will certainly be a very important area in the near future.

## CONCLUSIONS

The TMUX can be seen only as an interim device which is unlikely to be used after the analogue network has disappeared. Until then, it is one of several possible options for network planners for bringing about the transition from an analogue to a digital network. Telecommunication administrations world-wide are beginning to investigate their possible use both for inland and international networks. As they are used more frequently, so practical experience is gained and more uses for them are identified. The TMUX appears to be playing an increasingly important role in this transition and BTI has identified two areas where they are to be used. A large and expanding international market has grown from the origin concept in 1970.

## ACKNOWLEDGEMENTS

The author would very much like to thank STC, Basildon, who supplied the equipment for the trial, for allowing test results on the equipment to be used in this article and for their advice on technical aspects.

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### Biography

Andy Dick joined the International Lines Executive of BTI in 1982 after gaining an honours degree in Electrical Engineering at Dundee University. Since then, he has been concerned with transmission equipment in BTI's international repeater stations and, in particular, with transmultiplex equipment.

# Supergroup and Hypergroup Codecs

M. J. ANDREWS, B.SC., A.M.I.E.E.†

UDC 621.38.037.33 : 621.38.037.37

*Until now, analogue techniques have been used to develop British Telecom's transmission network. During the next decade, the change-over to an all-digital trunk transmission network will take place, and in the transition period it will be necessary to transmit analogue-multiplexed signals over digital transmission paths. The equipments described below provide this facility at two levels in the analogue multiplexing hierarchy.*

## INTRODUCTION

The CCITT* standardised hierarchies used in British Telecom's (BT's) network for analogue and digital multiplexing and transmission are shown in Table 1. Within the next decade, BT's network will be transformed from analogue to digital working with all trunk transmission being over digital line systems, and will use either coaxial cable, optical-fibre cable or microwave radio. During this period of evolution, a mixture of equipment using both analogue and digital techniques will coexist, and situations will arise in which the use of digital line transmission systems to convey wideband traffic originating from an analogue source will be advantageous. The use of the relevant frequency-division multiplexing (FDM) channel (CTE), group (GTE) and supergroup translating equipment (STE) to translate the analogue signals back to baseband, and the appropriate digital time-division multiplexing (TDM) equipment to remultiplex it would be possible, but such an approach would represent an uneconomic use of equipment and would present operational and practical difficulties.

Alternatively, analogue-to-digital (A/D) conversion and multiplexing techniques, analogous in certain respects to primary pulse-code modulation (PCM) encoding, can be applied directly to the wideband analogue signal. The supergroup and the hypergroup codecs make use of this technique.

The Equipment Supergroup Codec 2000 translates a basic analogue supergroup into a digital equivalent at 8·448 Mbit/s, and provides the complementary translation process. The Equipment Hypergroup Codec 6000 translates a basic analogue hypergroup into a digital equivalent at 68·736 Mbit/s, and provides the complementary translation process. (This latter bit rate is not yet an international standard, but is multiplexed on a conventional 34/140 Mbit/s digital multiplex (EDM 6002) by means of a tributary card which replaces two standard 34·368 Mbit/s inputs.) Whilst the techniques used may, as has already been indicated, bear some relationship to those used in the primary PCM encoder, in practice, the wider analogue bandwidths and higher digital line rates result in significantly different designs of equipment.

† Trunk Services, British Telecom National Networks

* CCITT—International Telegraph and Telephone Consultative Committee

**TABLE 1**
**Analogue and Digital Hierarchies**

| Analogue (kHz) | Voice Channels | Digital (kbit/s) |
|---|---|---|
| 4 (channel) | 1 | 64 |
| 60–108 (group) | 12 | |
| | 30 | 2 048 |
| 312–552 (supergroup) | 60 | |
| | 120 | 8 448 |
| | 480 | 34 368 |
| 312–4028 (hypergroup) | 900 | |
| | 1920 | 139 264 |

## PERFORMANCE CONSIDERATIONS

For a theoretical review of FDM/TDM encoding techniques, see Reference 1.

Two main sources of impairment, noise and jitter, are significant in relation to the operating performance of the codecs; both are explored below.

### Noise

#### *General*

In a codec, the main mechanism for noise generation is the coding/decoding process, the contribution from the analogue stages is small, although not negligible, in comparison. Sampling-process noise can itself be subdivided into quantising distortion and peak-clipping noise. Quantising distortion arises from the finite number of coding steps available to digitise the analogue input signal and synthesise the analogue output from the digital encode; the higher the resolution of the process—that is, the more coding steps—the lower the noise. Peak-clipping noise results when the amplitude of the input signal exceeds the level which produces the maximum peak digital code. Such amplitudes are, therefore, coded at a fixed level, and clipping is introduced when the analogue output is reconstituted. The degree of clipping depends on the operating range of the coder and the amplitude distribution of the input signal. The overall noise performance of the coder is thus related to these two phenomena, and depends heavily on the relationship between sampling resolution, coder operating range and signal amplitude distribution.

Fig. 1 relates these parameters for the hypergroup codec, assuming a fully-loaded hypergroup input, approximately

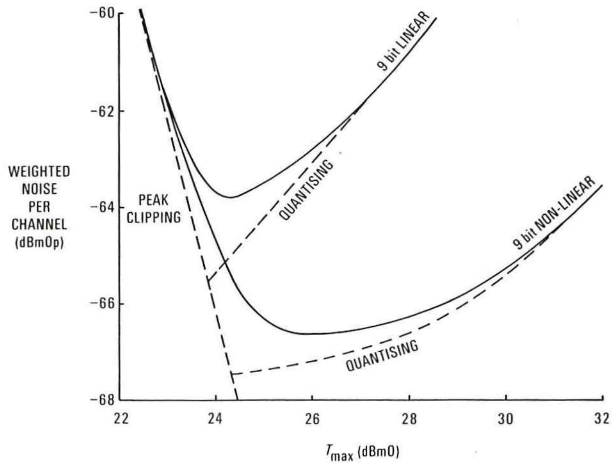



FIG. 1—Hypergroup codec, theoretical noise performance

equivalent to Gaussian noise. $T_{max}$ is the RMS value of a sine wave having peaks that just activate the positive and negative peak digital codes; it thus equates to the coders operating range. For a sine wave of RMS amplitude equal to $T_{max}$, no peak clipping would be expected. However, for the real signals encountered from traffic, where the amplitude probability distribution is less well defined, the choice of $T_{max}$ has an important bearing on clipping-noise generation. The 9 bit/sample resolution is constrained by the available line transmission rate, but the significant reduction in channel noise power resulting from non-linear coding is clear.

The coding law chosen, see Fig. 2, gives an effective 10 bit resolution to signals up to 1/4 full scale, 9 bit resolution for 1/4 to 1/2 full scale and 8 bit resolution for the remainder, and is made possible by the limited probability of higher amplitudes in the Gaussian distribution, relative to the lower amplitudes.

A relationship similar to that shown in Fig. 1 applies to the supergroup codec, where 12 bit linear coding produces an acceptable noise performance at a $T_{max}$ value of +21·5 dBm0. A fully loaded supergroup equates to a noise equivalent somewhere between that of a single channel and the Gaussian distribution appropriate to a hypergroup.

In practice, the noise performance achieved, measured in terms of psophometrically-weighted noise per channel, is better than 150 pW0p for the supergroup codec and better than 700 pW0p for the hypergroup codec, taking into account the unavoidable departures from the theoretical consequent upon the actual coder/decoder realisation, and including the additional noise contribution from the various analogue stages necessary in the codecs. Thus, the noise performance meets the limits that are expected to be defined internationally.

### *Noise Performance Under Light Loading*

When lightly loaded, the codecs exhibit an interesting performance characteristic that results in the noise deviating from a nominally uniform spectral distribution and, instead, bunching around certain frequencies, so that peaks in excess of the expected level are created.

The quantising-distortion noise waveform approximates to a sawtooth of amplitude equal to the spacing between quantising levels, and contains harmonics of the signal frequency[2]. As a result of the sampling process, intermodulation between these harmonics and the sampling frequency and its harmonics produces components within the signal passband. Where a simple relationship exists between the sampling frequency and the signal frequency, these intermodulation products congregate around sub-harmonics of the sampling frequency and give rise to local noise peaks.

In practice, the presence of a few FDM system pilots and carrier leaks provides sufficient loading to obviate the effect, and no problems are expected in operational service.

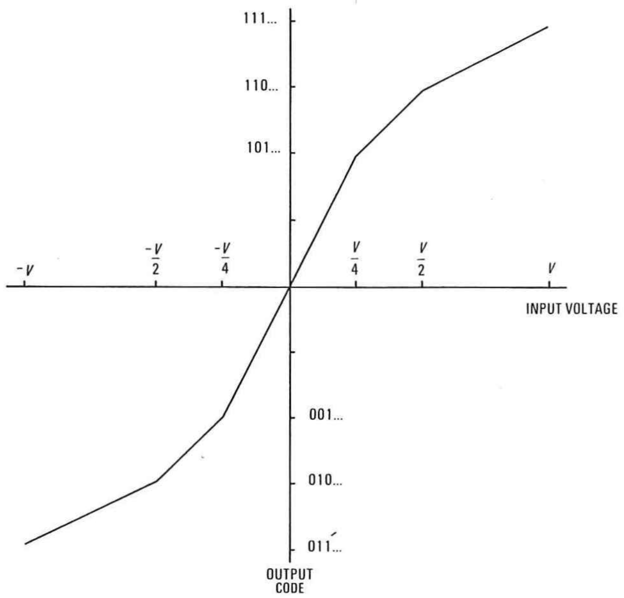



FIG. 2—Hypergroup codec, non-linear coding law

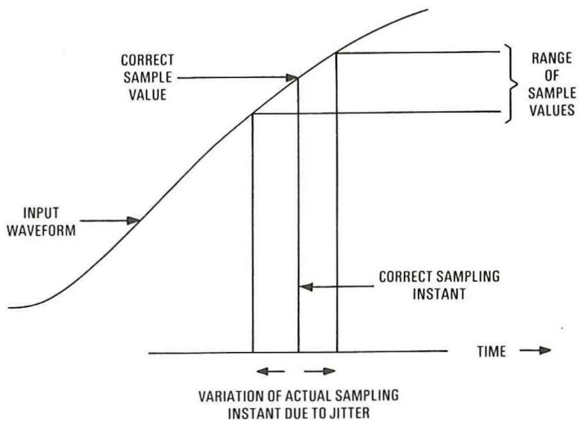



FIG. 3—Aperture uncertainty

## Jitter

### *Analogue-to-Digital Conversion*

An A/D converter operates by sampling the input analogue signal at specific time instants and producing digital encodes of the signal at these instants. If there is jitter on the clock providing the sampling signal, the analogue signal level as sampled deviates from that expected under ideal conditions (that is, no jitter present). In practice, it is difficult to eliminate jitter totally, and the effect of sampling with a jittered clock—known as *aperture uncertainty*—is to introduce noise into the coding/decoding process. Fig. 3 illustrates the mechanism. To keep the noise introduced from this source within acceptable limits, the sampling frequency jitter on the supergroup codec must be less than 70 ps, and on the hypergroup codec less than 40 ps, and this demands particularly careful attention to the design and layout of the equipment.

### *Digital-to-Analogue Conversion*

Any jitter introduced onto the signal at the digital input to the receive codec by the digital line system manifests itself as phase-noise sidebands on the wanted decoded analogue signal and, hence, increases the noise level. Conventional techniques are employed to reduce these effects; for example, appropriate choice of loop bandwidth in the timing-recovery circuits.

The practical realisation of the two codec designs are now discussed in more detail.

## EQUIPMENT SUPERGROUP CODEC 2000

Fig. 4 shows a simplified block diagram of the supergroup codec. As much of the design uses conventional circuit techniques, only a brief description of these is given, emphasis being laid on the more unusual aspects of the design. The input filter provides a degree of rejection of out-of-band frequencies generated in preceeding analogue multiplexing equipment (either GTE or STE), and equalisation to correct any amplitude/frequency distortion introduced by interconnecting cables. The filtered and equalised analogue signal is then sampled at 576 kHz in the A/D converter, which produces 12 bit parallel linearly-encoded sample words. These are converted to 14 bit words in the multiplexer, by the addition of two extra bits per sample

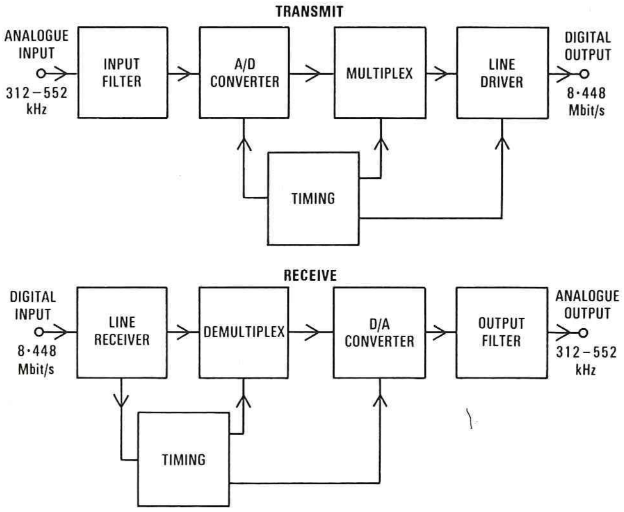



FIG. 4—Supergroup codec, simplified block diagram

(service digits). Blocks of 12 of these 14 bit words are assembled with an 8 bit frame-alignment word (11101000) to form a 176 bit frame, at a frame repetition rate of 48 kHz.

In the line driver, the resultant 8·448 Mbit/s serial binary bit stream is converted to the required HDB3 (high-density bipolar 3) line code for connection to other digital line or multiplexing equipment.

In the receive direction, the equivalent reverse operations are performed to recover the analogue signal from the incoming 8·448 Mbit/s HDB3 encoded digital input signal.

## Analogue-to-Digital Converter

The A/D digital converter, see Fig. 5, is the heart of the codec and employs a two-stage conversion technique to achieve the required digital accuracy at the necessary sampling rate.

The buffered analogue input is sampled and the sample level stored in Sample and Hold No. 1. A/D Converter No. 1, an 8 bit successive-approximation register (SAR) device, (see Appendix) produces a digital encode of the sample level; the six most-significant bits are passed to the logic combining stage and to the most-significant input bits of the 12 bit D/A converter. The output of this device is subtracted from the original input sample value, now transferred to Sample and Hold No. 2, and the difference signal, which is the residue left after the coarse first-stage conversion, is encoded in a second 8 bit SAR A/D converter, whose seven most significant bits also pass to the logic combining stage where, after error correction, the 12 bit composite sample word is produced.

The two sample-and-hold stages allow the overall sampling rate of 576 kHz to be maintained with the two-stage conversion technique; Sample and Hold No. 1 and A/D Converter No. 1 begin to process a new input sample while A/D Converter No. 2 is still processing the previous one. The sample-and-hold modules also keep the input signal level to the SAR A/D converters constant during their conversion cycle.

The use of the various A/D and D/A converters over limited sections of their available dynamic ranges increases the overall accuracy of the conversion processes, and the additional bit sent to the logic combining stage allows any overlap errors created in the two-stage conversion to be corrected in the final 12 bit sample.

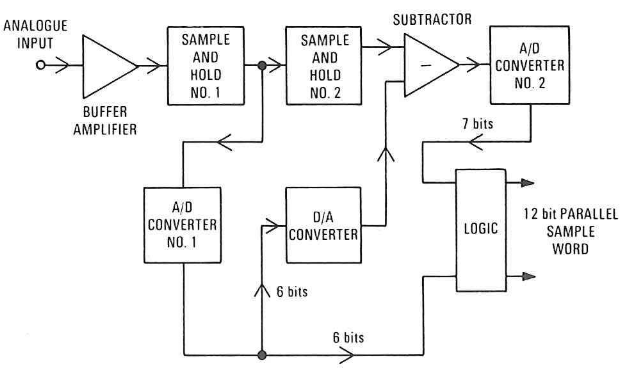



FIG. 5—Supergroup codec, analogue-to-digital converter

## Digital-to-Analogue Converter

The 12 bit sample words recovered from the input digital stream via the line receiver and demultiplexer are applied to a D/A converter, which in practice is a single package commercial hybrid device that produces a current output corresponding to the value of the digital input word. Because of the choice of sampling frequency, however, it is not sufficient to apply this current directly to a simple current-to-voltage converter to produce the analogue output signal, and an examination of the sampling process used reveals the reason.

In most sampling systems, the sampling frequency is chosen to be at least twice the highest signal frequency. One reason for this is to ensure that sidebands produced in the sampling process do not have frequency components which could interfere with the signal being sampled, a phenomenon known as *aliasing*, see Fig. 6. Whilst the 2 × frequency convention applies to input signals whose frequency spectrum starts at close to zero, the supergroup codec has an input frequency spectrum which begins at 312 kHz and extends to 552 kHz. The gap between 0 and 312 kHz is thus wider than the signal spectrum (240 kHz) and, by careful choice of sampling frequency, the relevant sampling sideband can be arranged to fall neatly into this slot, and so avoid aliasing and restricting the required sampling frequency to less than twice 552 kHz (see Fig. 7).

In practice, other constraints, for example, availability of suitable encoders and the allowable digital line bit rate, dictate that such an approach is essential, and a sampling frequency of 576 kHz is found to satisfy all the conflicting requirements.

However, the choice of 576 kHz as sampling frequency introduces a problem when it comes to decoding back to analogue. The level/frequency performance of a general

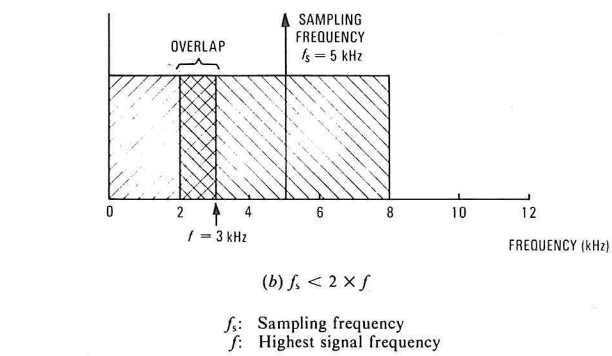



FIG. 6—Choice of sampling frequency

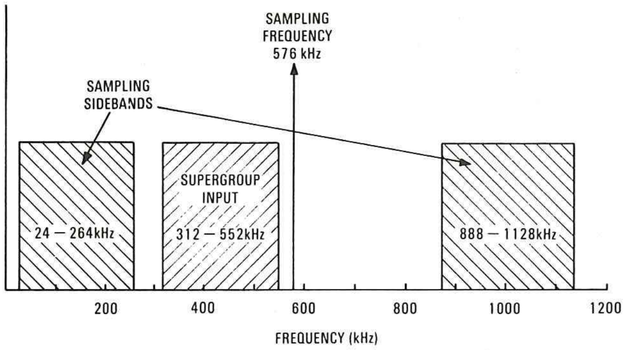



FIG. 7—Supergroup codec, sampling spectrum

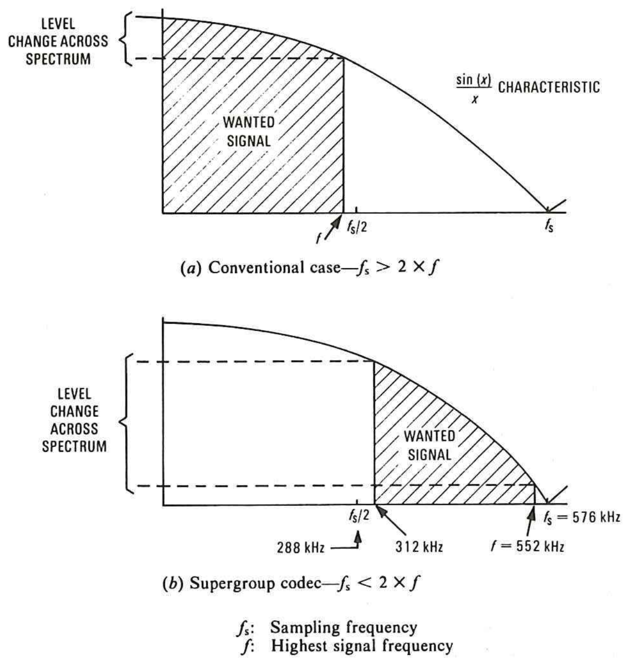


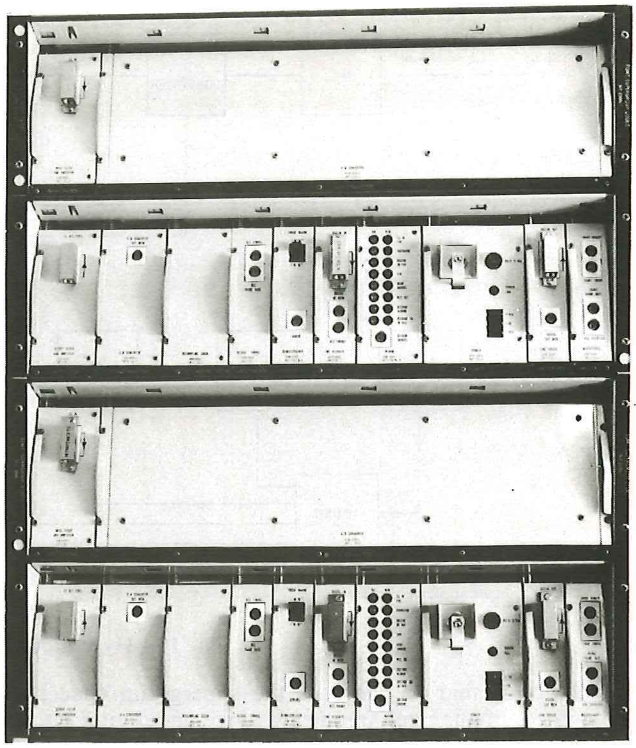

FIG. 8—Frequency distortion introduced by sampling process

decoding process is shown in Fig. 8. In the case of the conventional method of using sampling at twice the highest frequency (Fig. 8(*a*)), the wanted signal in the range 0 to $f_s/2$ occupies a portion of the response curve where the amplitude/frequency distortion is low, and can easily be corrected by conventional equalisation techniques. However, for the supergroup codec, the wanted portion of the spectrum lies on a part of the response where the amplitude/frequency distortion is large (Fig. 8(*b*)); the change in level over the range 312–552 kHz exceeds 22 dB, and presents particular problems for the design of suitable equalisation circuitry.

To ease the requirements of the subsequent equaliser, the supergroup codec uses a resampling switch between the D/A converter output and the current-to-voltage converter. The resampler acts as a gate to the D/A converter output signal and allows either the signal or a zero level to be passed to the following circuitry. With a mark:space ratio of 5:6, the effect of this finite width sampling is to impose an additional layer of frequency distortion onto the D/A output, and reduce the overall level/frequency variation to approximately 2 dB over the wanted spectrum, but at the expense of absolute level. The remaining equalisation necessary, now much simpler to implement, together with additional gain to bring the output level to that required, is provided in the output filter unit.

## Alarms

Comprehensive alarm facilities are provided to monitor the performance of the codec. In the transmit direction, *loss of supergroup input* and *A/D converter overload* are indicated, detected from the activity of various code levels at the A/D converter output. In the receive direction, *loss of digital input*, *loss of alignment*, *high errors* and *AIS* are indicated. In addition, two alarms are transmitted from the far-end codec in the service digits already mentioned; these are *loss of distant supergroup input* and *distant alarm*, the latter activated by a loss of digital input or loss of alignment in the far-end receive side. Means are also provided, at test ports, to monitor the error performance at the far-end codec, this information being sent in coded format in further service digits. Test ports used in conjunction with the alarm indications allow faulty cards to be identified.



*Note*: Two equipments, each equipment occupies two shelves

FIG. 9—Supergroup codec

## Mechanical Construction

The supergroup codec occupies two shelves of 62-type equipment practice[3] (Fig. 9) and operates from nominal 28 or 48 V DC station power supplies using a single high-efficiency switch-mode power unit.

Because of the mixture of low-level wideband analogue signals and high-level high-frequency digital signals found in the equipment, comprehensive screening of individual cards is necessary, in particular on the A/D converter, where the individual circuit boards are housed in compartments formed in a solid aluminium block. As a result, the A/D converter occupies almost one complete shelf.

Other than connections to the station power and alarm system and to the analogue and digital input and output ports, the supergroup codec is totally self-contained, generating all its timing signals in the transmit direction from a crystal-controlled oscillator running at 50·688 MHz, and deriving its timing signals in the receive direction from the 8·448 Mbit/s digital input signal.

## EQUIPMENT HYPERGROUP CODEC 6000

Whilst the basic principles employed in the hypergroup

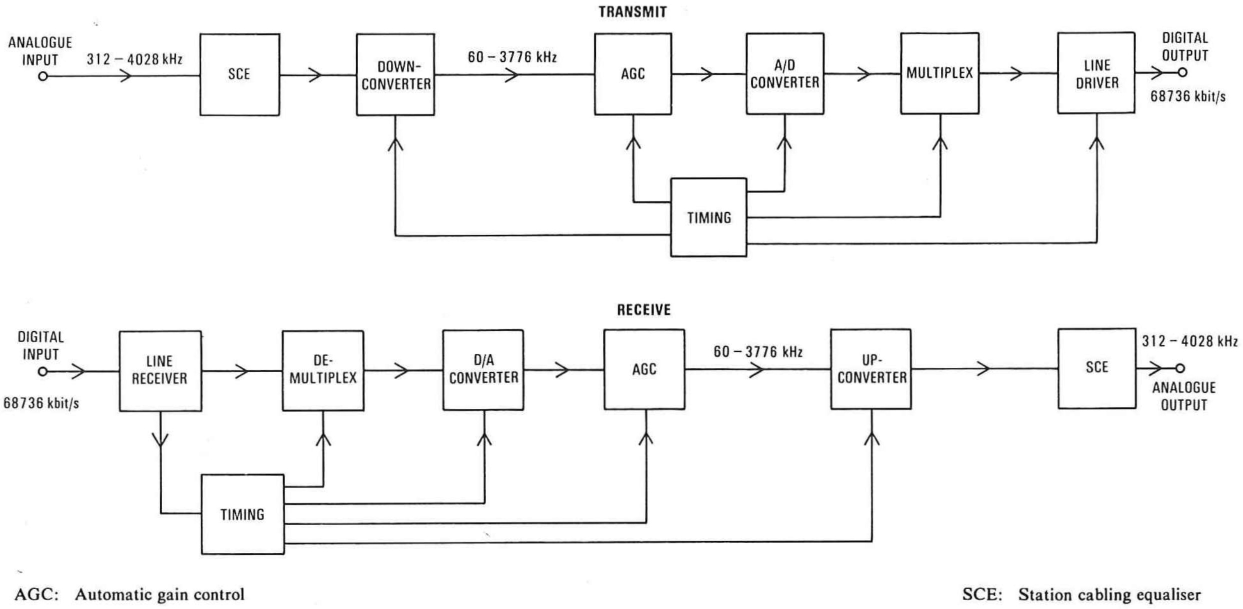



FIG. 10—Hypergroup codec, simplified block diagram

codec are similar to those of the supergroup codec, the differences in analogue frequency range and digital bit rate result in a totally different practical equipment.

Fig. 10 shows a simplified block diagram of the equipment, from which it can be seen that several additional functional blocks have been introduced.

The choice of key parameters for a translation between the analogue and digital domains is affected by a number of factors. As has already been explained, with any analogue-to-digital-to-analogue conversion, noise, known as *quantising noise*, is introduced as a consequence of the finite number of quantising levels available. Where digital samples are to be transmitted over a line system, the allowable bit rate, the logistics of frame alignment loss and recovery times, and ease of generation of particular frequencies have a bearing on the choice of sampling frequency, frame rate etc. For the hypergroup codec, the combination of these factors leads to the use of 9 bits per sample, non-linear encoding at a sampling rate of 7608 kHz and a frame rate of 24 kHz.

With an input signal frequency range of 312–4028 kHz, there is no scope for sampling at less than half the highest frequency, but to avoid aliasing with the chosen sample rate of 7608 kHz, the maximum signal frequency cannot exceed 3804 kHz. It would be possible to limit the input frequency to this maximum by excluding the top supergroup, and thus limit transmission to a 14-supergroup assembly, but this would introduce operational difficulties. To accommodate a full 15-supergroup assembly, the input signal is frequency translated downwards by 252 kHz to occupy the range 60–3776 kHz prior to encoding, and retranslated back to 312–4028 kHz subsequent to decoding; this function being carried out in the down- and up-translators, respectively.

The noise performance described earlier assumed a fully loaded hypergroup. In practice, the actual loading in a hypergroup can vary over wide limits, from a few channels upwards, depending on time of day, type of traffic etc, and even the non-linear coding law is insufficient to ensure acceptable performance under all extremes. An automatic gain control (AGC) circuit is therefore used to ensure that the input signal to the A/D converter occupies, as far as possible, the optimum operating range of this converter to minimise the noise contribution.

## Translation Units

The down-translator uses two-stage modulation, with carriers at 10 MHz and 9·748 MHz, to convert the input signals to the frequency range 60–3776 kHz.

The up-translator performs the complementary translation on the output of the D/A converter to return the signals to the range 312–4028 kHz, again using two-stage modulation and with the same carrier frequencies. To ensure zero overall frequency shift through the two translation stages, the down-translator carriers are frequency locked to the transmit frame rate, whilst those in the up-translator are frequency locked to the digital input signal received by the codec, and hence to the same transmitted frame rate.

## Automatic Gain Control

The hypergroup codec analogue input signal consists of a signal assembled from 15 supergroups, each comprising 5 groups of 12 channels, 900 channels or their equivalent in total. Depending on time of day, type of traffic and the originating source of the circuit, the activity within this block varies over a significant range, anything from a few channels to complete loading. Consequently, the dynamic range of the signal applied to the A/D converter is large.

The use of non-linear encoding improves the overall performance of the codec relative to linear encoding, but further enhancement is desirable. An AGC circuit is therefore used to maintain the signal offered to the A/D converter at a level which occupies its coding range to best advantage. By using such means, an improvement of up to 8 dB over the non-AGC condition can be achieved for low-level signals, and up to 4 dB for high levels.

The AGC circuit is shown in Fig. 11. A gain-controlled amplifier at the input to the A/D converter is driven by a control signal derived from digital detection of the activity of the converter output. The gain applied to the input signal is therefore varied in accordance with its level. To provide a control signal for the receive-end codec, a fixed-level 12 kHz tone is added to the translated hypergroup at the input to the gain-controlled amplifier. The level of this control signal at the output of this amplifier, and hence at the input to the A/D converter, is thus dependent on the gain setting of the AGC amplifier.

At the receive end, after decoding, the level of the 12 kHz signal is monitored and compared to a reference, and the difference is used to control a second gain-controlled amplifier so as to reverse the effect of the AGC amplifier at the transmit end. The 12 kHz pilot is filtered from the wanted signal at this stage.

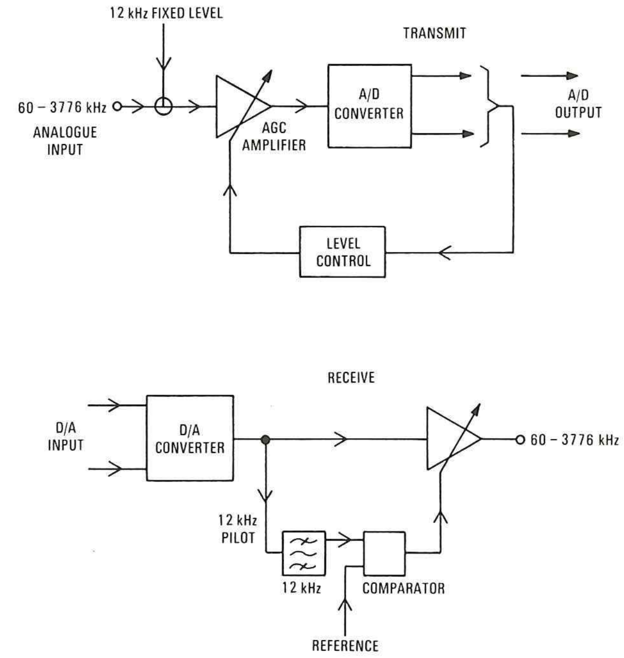



FIG. 11—Hypergroup codec, automatic gain control

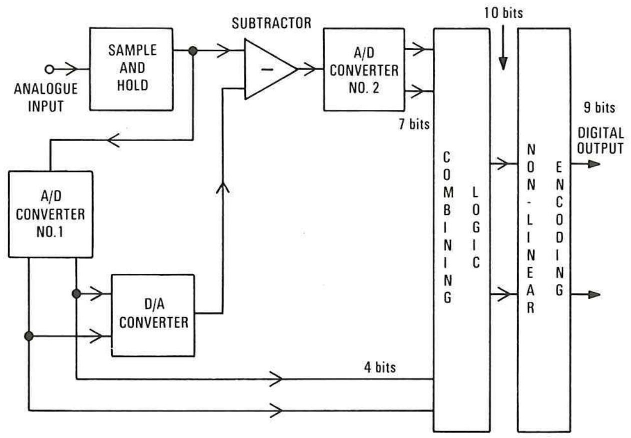



FIG. 12—Hypergroup codec, analogue-to-digital converter

### Analogue-to-Digital Converter

The 9 bit non-linear encoding is achieved by digitally transforming the linear 10 bit output of the A/D converter. As with the supergroup codec, a two-stage encoder is used. However, only a single sample-and-hold circuit is required as the individual 4 and 8 bit A/D converters are high-speed parallel, or flash, converters (see Appendix). Fig. 12 shows the basic arrangement of the encoder. Similar comments to those mentioned in relation to the supergroup codec apply to the use of converters over limited dynamic ranges and the overlap correction bit.

### Multiplexer and Line Driver

The hypergroup codec frame consists of a 9 bit frame alignment word, two bits for service digits and 317 sample words, each of 9 bits, assembled in the multiplexer unit.

The line code used for the digital interfaces on the hypergroup codec is coded mark inversion (CMI)[4]. Encoding of the serial digital multiplexed signal to this required line code is also performed in the multiplexer, the reverse process taking place in the receive side of the codec in the demultiplexer.

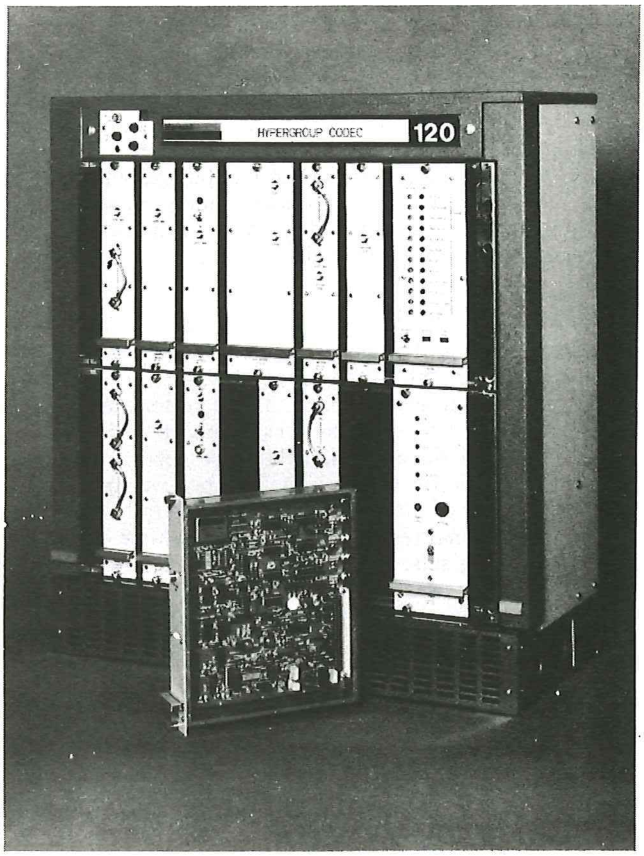



FIG. 13—Hypergroup codec

### Digital-to-Analogue Converter

The 9 bit samples received from the demultiplexer are converted to 10 bit linear by a logic transformation. The resulting 10 bit sample words are applied to a 10 bit D/A converter, the output of which is sampled in a sample-and-hold circuit to eliminate switching transients caused during the conversion process before being passed to the receive AGC unit.

### Alarms

Equivalent alarms to those found on the supergroup codec are provided, together with additional monitors for the operation of the translation and AGC stages.

### Mechanical Construction

The hypergroup codec is constructed in TEP-1(E) construction practice[5] and occupies two 8 VU shelves, see Fig. 13. To avoid interference problems associated with the mix of low-level wideband analogue and high-level high-speed digital signals, comprehensive screening arrangements are used for all low-level stages. Operation is possible from nominal 28 or 48 V DC station supplies using a single power converter, and all clock signals required by the equipment are derived internally from a 68 736 kHz crystal-controlled oscillator in the transmit direction, and from the line signal in the receive direction.

## OPERATIONAL ASPECTS

### Codec Application

The digital signals produced and accepted by the codecs are compatible in terms of amplitude and line code with the

respective digital multiplexing or line terminating equipment interfaces. However, the frame structures of the codec digital signals are unique to these equipments, and thus require a codec at each end of a digital route, although such routes may include digital multiplexing as well as transmission. The digital output of a codec cannot, therefore, be demultiplexed to channel level via conventional digital multiplexing equipment. The codecs are, however, uniquely suitable for carrying non-channelised analogue traffic, such as wideband data and sound programme circuits.

### Codec Maintenance

In-service testing of the operation of the codecs is possible by using Tester 281 for the supergroup codec and Tester 282 for the hypergroup codec. These testers monitor the frame-alignment words of digital systems operating at 8 and 34 Mbit/s, and 68 and 140 Mbit/s respectively, and are switch selectable for use with their respective codecs.

Maintenance of faulty equipment is generally possible by using conventional test equipment; for example, oscillators, level measuring sets, frequency counters etc. However, the A/D converters used in both codecs do not lend themselves so readily to simple testing, and more sophisticated techniques are being explored.

## CONCLUSIONS

The supergroup and hypergroup codecs will assist in the smooth change-over from an analogue to a digital transmission network. This article has indicated the degree of sophistication employed in their design and their ease of application in the working environment.


## ACKNOWLEDGEMENTS

The supergroup and hypergroup codecs were developed by Marconi Communications Systems Ltd. under British Telecom development contracts.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### Biography

Mike Andrews joined the then BPO as an Executive Engineer in 1971 after graduating from Durham University, and worked initially on power supplies and general standards for transmission equipment. After promotion in 1979, he worked in the FDM development group where his work included the codec developments. Currently, he is head of the transmission test equipment development group in BT National Networks and represents BT interests on several CEPT and CCITT committees.

## APPENDIX

## ANALOGUE-TO-DIGITAL CONVERTERS

Two main types of A/D converter are used in the codecs. In the supergroup codec, the successive approximation register converter, and in the hypergroup codec the parallel, or flash, converter.

### Successive Approximation Register Converter

With reference to Fig. 14, the input signal to be digitised is applied to a comparator. The second input to the comparator derives from a D/A converter whose digital inputs are driven by a counter, itself driven by a clock.

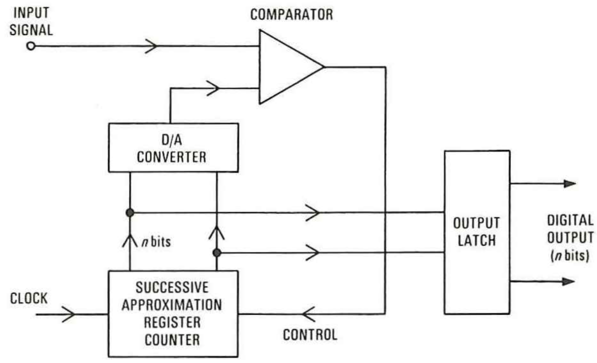



FIG. 14—Successive approximation register A/D converter

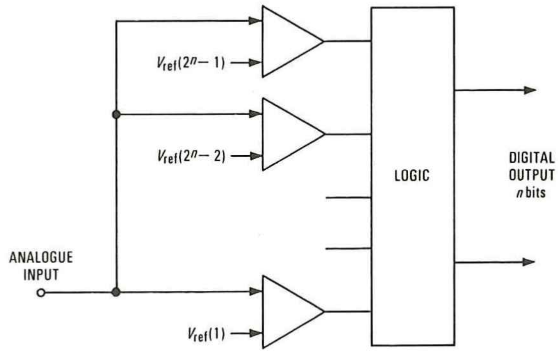



FIG. 15—Parallel, or flash, converter

The counter activates one bit of the D/A at a time starting with the most significant bit (that is, half full scale). The comparator output indicates which of its two input signals is the larger, and this information is fed back to the counter. As each bit of the D/A converter is activated, the comparator indicates whether the resultant generated signal is less than or greater than the analogue input signal; if less, the bit stays on, if greater, the bit is turned off. The counter then activates the next most significant bit of the D/A converter and the process continues until all bits have been tried, at which time the bit pattern at the output of the counter represents the digital encoded value of the analogue input signal, within the accuracy allowed by the number of bits available. At this point the cycle repeats. The converter, therefore, approaches the final value one bit at a time, starting with one half full scale and with each successive change being one half of the previous one; only when all bits have been tried is the output signal valid.

The output latch is enabled at the end of each conversion cycle to ensure that only valid outputs are presented. Whilst the converter is going through its routine, the input signal must be held constant. The sample-and-hold circuits described in the text provide this facility. As each conversion is completed, the sample-and-hold circuit acquires a new input signal ready for the next cycle.

### Parallel, or Flash, Converters

In the parallel A/D converter, see Fig. 15 the input signal is presented simultaneously to a bank of comparators, each set to operate at a slightly different level, and the required digital code is produced from the output states of the comparators. The parallel converter is thus very fast, but requires as many comparators as there are code values. Therefore, as resolution increases so does circuit complexity, and the required accuracies of the comparator reference voltages and stability of the comparator trigger points become increasingly more difficult to maintain. The main use of parallel converters, therefore, tends to be in high-speed lower-resolution applications.

# A 2048 kbit/s Data-in-Voice Modem

N. HARRISON, B.SC., A.M.I.E.E.†

UDC 621.38.037.37 : 621.38.037.33

*While the change-over from an analogue to a digital trunk network takes place, there may be occasions when it could be considered advantageous to have the capability of routeing digital signals via analogue bearers. These advantages could be seen, perhaps, in terms of short-term costings and speed of deployment.*

*If a bit-sequence-independent digital transmission capability is required, then modems must be used. This article examines a particular modem* which has been designed to convey a 2048 kbit/s digital signal over two contiguous supergroups within an analogue hypergroup assembly.*

## INTRODUCTION

Towards the end of 1982, British Telecom (BT) National Networks obtained four 2048 kbit/s data-in-voice (DIV) modems for evaluation. These particular modems allow the transmission of a bit-sequence-independent 2048 kbit/s digital signal within the bandwidth usually allocated to supergroups 8 and 9 of an analogue hypergroup assembly.

These modems are known as *data-in-voice* modems because they use a part of the hypergroup spectrum normally associated with frequency-division multiplexed (FDM) speech channels. (Data-over-voice (DOV) modems also exist, and these use the spectrum above that normally associated with FDM speech channels.) The CCITT has been considering these types of equipment and has formulated Recommendation G941[1] to describe the versions that handle the higher bit rates.

The modems have been carefully examined and subjected to both laboratory and field evaluations. This article describes the findings and observations from this work, dealing with the following aspects:

(*a*) the interconnection of the modem between the digital and analogue networks,

(*b*) the signal processing techniques used in the modem,

(*c*) the installation and operational considerations (for example, alarms, equipment practice, etc.),

(*d*) the results of the laboratory and field evaluations, and

(*e*) the fundamental limitations of an analogue network when carrying digital data.

In closing, the article makes reference to future work and summarises the more important findings.

## INTERCONNECTION BETWEEN THE DIGITAL AND ANALOGUE NETWORKS

The modem converts an HDB3‡ encoded 2048 kbit/s digital signal to an analogue signal that lies within the bandwidth normally allocated to supergroups 8 and 9 of a hypergroup assembly; that is, from 1804–2292 kHz. To gain access to this portion of the hypergroup spectrum, a wideband input/output facility at the supergroup translating equipment (STE) is required. To provide this facility, the STE requires the fitting of an optional card set, known as an *addition supergroup (ADD SG) equipment*, which occupies one shelf of the STE. However, although most STEs have the capability to accept the ADD SG equipment, it is not usually found fitted except where specifically requested when the STE was originally ordered.

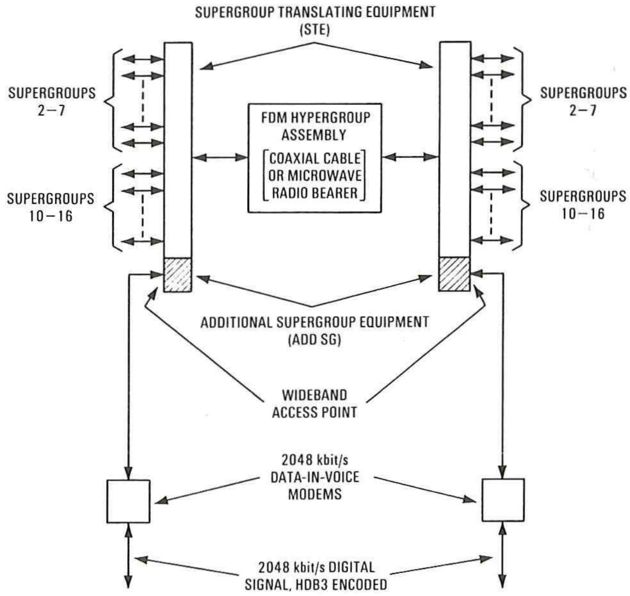



FIG. 1—Interconnection of the modem to the digital and analogue networks

Fig. 1 shows the modem connected to the STE via the ADD SG equipment.

As mentioned previously, the ADD SG equipment is a wideband access point with respect to the hypergroup spectrum. Thus, if the analogue signal from the modem is to be through-connected into a different hypergroup, a special through connect filter is required immediately after the ADD SG output point.

The modem manufacturer, having considered this application, has produced a filter that possesses the required band-pass characteristics over supergroups 8 and 9. However, it is recommended that not more than one of these filters be used between any pair of modems because of the effects of group-delay distortion.

The analogue signal emanating from the modem represents a loading of about +6 dBm0 into the FDM network. This is comparable with the equivalent FDM speech channel loading of two supergroups; that is, 120 speech channels.

† Trunk Services, British Telecom National Networks

* The modem referred to in this article is proprietary equipment manufactured by Fujitsu Ltd. of Tokyo, Japan

‡ High-density bipolar type-3 interface code

## MODEM OPERATION

Fig. 2 shows a block diagram of the various subsystems that comprise the modem. With reference to this diagram, the signal path is now traced through each of these subsystems, and each signal processing operation encountered is briefly explained.

### Digital-to-Analogue Signal Processing

#### *Coder Subsystem*

The HDB3 encoded 2048 kbit/s digital signal is initially converted to a unipolar (that is, binary) format in the bipolar-to-unipolar converter (B–U CONV) unit. This unit also extracts a 2048 kHz timing signal from the digital input for use as the master timing signal for the remainder of the transmit circuitry. The binary signal is then scrambled in a self-synchronising 22-stage scrambler (SCR), having the characteristic polynomial

$$f(x) = x^{22} + x + 1;$$

that is, the binary input is modulo-2 added to 22- and 1-unit delay versions of itself to produce the scrambled binary output.

The scrambling operation has the following objectives:

(*a*) to randomise the binary signal and thus help prevent 'strong' localised spectral components in the final analogue signal,

(*b*) to minimise the low-frequency components of the binary signal (produced, say, as a result of a long sequence of binary ones), and

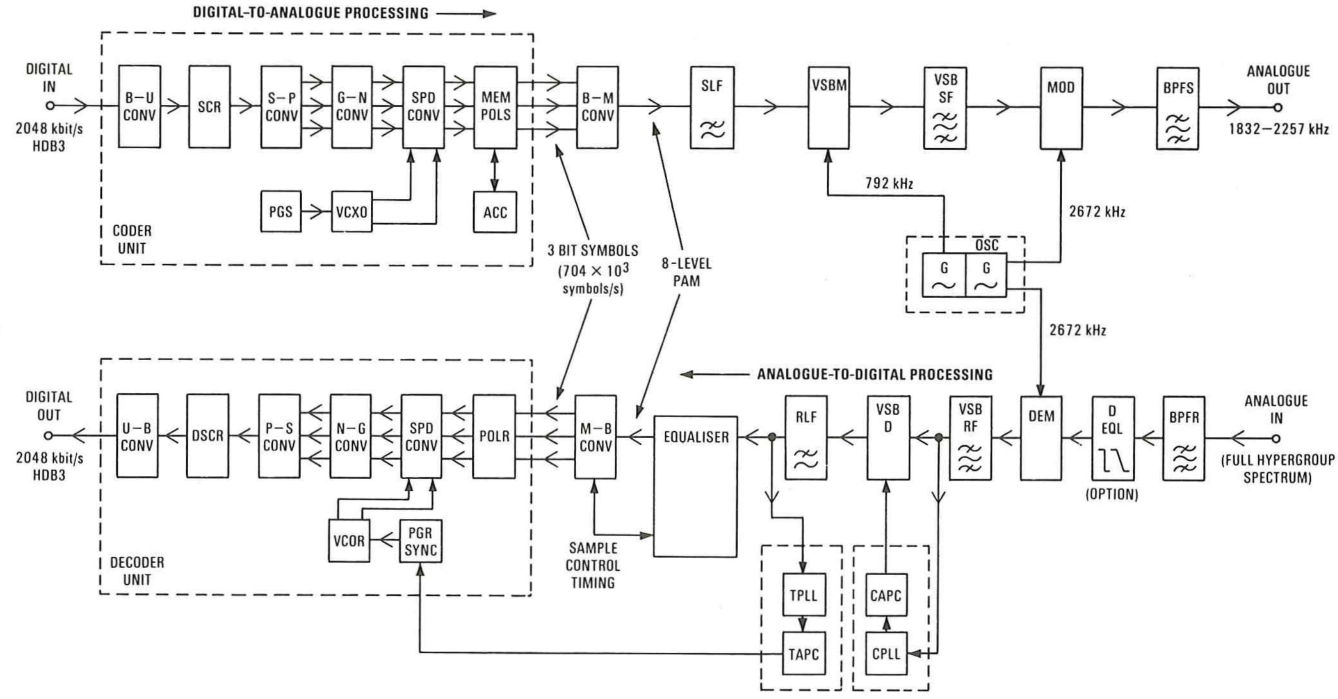
(*c*) to prevent 'strong' correlation of adjacent multilevel codes that are to be subsequently produced.

It is worth noting here that the corresponding descrambling operation, performed in the receiving modem, introduces an error multiplication factor of three; that is, a single binary error entering the descrambler results in three binary errors out of the descrambler. This is caused by the two feedback taps of the descrambler circuit.

The scrambled binary signal is then converted, from the 2048 kbit/s serial format, into a 3-path parallel format in the serial-to-parallel converter (S–P CONV) unit. The output of this unit can be regarded as a 3 bit wide bus working at a 3 bit symbol rate of $682 \cdot 66' \times 10^3$ symbols per second (that is, $682 \cdot 66' \times 10^3 = 2048 \times 10^3/3$).

Each 3 bit symbol is then pre-coded from an assumed original Gray code format into a natural binary format. This process is performed in the Gray-to-natural-binary converter (G–N CONV) unit. The objective of this pre-coding operation is to minimise binary errors when the analogue signal is incorrectly decoded in the receiving modem due to noise corruption incurred during transmission. A detailed discussion of the pre-coding operation is given in the Appendix.

The next operation performed in the coder subsystem is feedback balanced coding[2] (FBC), which is implemented by the speed conversion (SPD CONV) unit, the memory and polarity send (MEM POL S) unit and the accumulator (ACC) unit. The FBC operation ensures that the subsequently produced 8-level pulse-amplitude modulated (PAM) signal has the following spectral characteristics:




ACC: Accumulators, long- and short-term
B–M CONV: Binary-to-multilevel converter (3 bit symbol to 8-level PAM)
B–U CONV: Bipolar (HDB3)-to-unipolar (binary) converter
BPFR: Band-pass filter, receive
BPFS: Band-pass filter, send
CAPC: Carrier (792 kHz) automatic phase control
CPLL: Carrier (792 kHz) phase-locked loop
DEM: Demodulator (2672 kHz)
D EQL: Optional (extra) group-delay equaliser
DSCR: Descrambler
EQUALISER: 41-tap transversal delay equaliser
G–N CONV: Gray-to-natural-binary converter
M–B CONV: Multilevel-to-binary converter (8-level PAM to 3 bit binary symbol
MEM POL S: Memory and frame polarity inverter (performs feedback balanced code operation)
MOD: Modulator (2672 kHz)
N–G CONV: Natural-binary-to-Gray-code converter
OSC: Oscillator unit
P–S CONV: Parallel-to-serial converter
PGR SYNC: Pulse generator from recovered timing (352 kHz)
PGS: Pulse generator, send
POL R: Frame polarity correction
RLF: Receive low-pass filter
S–P CONV: Serial-to-parallel converter
SCR: Scrambler ($f(x) = x^{22} + x + 1$)
SLF: Send low-pass filter
SPD CONV: Speed converter
TAPC: Timing (352 kHz) automatic phase control
TPLL: Timing (352 kHz) phase-locked loop
U–B CONV: Unipolar-to-bipolar converter
VCOR: Timing signal recovery
VCXO: Timing signal generator
VSB D: Vestigial sideband demodulator
VSBM: Vestigial sideband modulator
VSB RF: Vestigial sideband receive filter
VSB SF: Vestigial sideband send filter

FIG. 2—Block diagram of the modem subsystems

(*a*) very little energy in the low-frequency components, and

(*b*) a spectral null at 352 kHz to allow for the later insertion of a timing pilot at this frequency.

The particular FBC technique used in this modem is now briefly described.

The 3 bit symbol train is initially subject to a speed conversion process, which increases the 3 bit symbol rate from $682{\cdot}6\dot{6} \times 10^3$ symbols per second to $704 \times 10^3$ symbols per second; that is, an increase by the factor 33/32. This action is necessary to provide an extra 'time-slot' for the insertion of a framing symbol at a subsequent stage in the FBC process.

However, before the addition of the framing symbol can be considered, it is first necessary to divide the 3 bit symbol train into blocks of consecutive $32 \times 3$ bit symbols known as *frames*. Each frame is then further subdivided into ODD and EVEN sub-frames by taking alternate 3 bit symbol samples from the frame. Thus, each sub-frame consists of $16 \times 3$ bit symbols.

Each 3 bit symbol, in both the ODD and EVEN sub-frames, is given a weighted polarity value that is directly proportional to the natural binary magnitude departure from an imaginary zero-weighting mid-value point lying between the symbols 011 and 100; that is, levels 3 and 4, respectively, of the subsequently produced PAM signal.

The ACC unit compares the aggregate weight of the current ODD and EVEN sub-frames with an accumulated long-term running average that is kept for each. The objective is to keep both ODD and EVEN long-term running averages at or near a zero weighting. This is achieved by inverting (that is, complementing) the current ODD and/or EVEN sub-frame, in the MEM POL S unit, as required. When a sub-frame is inverted, the effect of this action is taken into account in the respective long-term running average.

So that the receiving modem can identify when a sub-frame has been inverted, and thus correct the process, a 3 bit framing symbol containing this information is added to each frame. The framing symbol is constrained to take on only one of four distinct values:

111 (level 7), meaning that no sub-frames have been inverted;

101 (level 5), meaning that the ODD sub-frame has been inverted;

010 (level 2), meaning that the EVEN sub-frame has been inverted; or

000 (level 0), meaning that both ODD and EVEN sub-frames have been inverted.

Note the following:

(*a*) The spacing between two adjacent allowable levels is at least two levels; this makes the framing symbol more robust against noise, which could cause it to be incorrectly decoded.

(*b*) The framing signal can be easily distinguished from the information symbols, over a long time period, since it can take on only four out of the eight possible levels; this feature of the signal is used as an aid to frame alignment.

(*c*) The addition of a 3 bit framing symbol has a minor effect on the overall weighting of the frame. However, this effect is allowed for by making the appropriate weight correction to the ODD and EVEN long-term running average accumulators as necessary.

There are still two remaining units in the coder subsystem to be discussed: the send pulse generator (PGS) and the speed conversion clock (VCXO). These units essentially act together to produce all the timing signals required by the transmit circuitry. They can also produce a 2048 kHz timing signal, to an accuracy of better than ±50 parts per million, in the event of loss of primary timing from the 2048 kbit/s digital input signal.

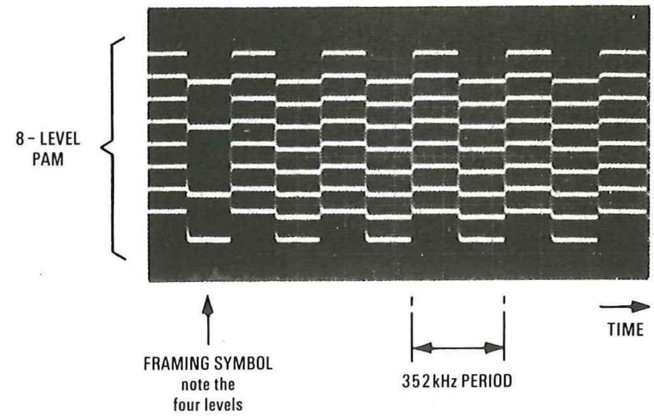


FIG. 3—Eight-level pulse-amplitude modulated signal, showing the 352 kHz square-wave modulation (timing pilot)

### *Binary-to-Multilevel Converter Subsystem*

The binary-to-multilevel converter (B–M CONV) subsystem produces an 8-level PAM signal by applying the 3 bit parallel binary symbols to a constant-current-drive ladder network; that is, a form of digital-to-analogue converter. The 352 kHz timing signal is also introduced by this unit, and has the effect of 'square-wave' modulating the 8-level PAM signal, as shown in Fig. 3. The modulation can be pictured as shifting each alternate 8-level PAM signal position up by one level.

### *Send Low-Pass Filter Subsystem*

The send low-pass filter (SLF) subsystem restricts the bandwidth of the PAM signal by using a raised-cosine low-pass filter characteristic[3] (which is a practical implementation of the Nyquist limit) defined, in the frequency domain, by:

$$P(f) = \begin{cases} \dfrac{1}{2B_T} & |f| < f_1 \\[2ex] \dfrac{1}{4B_T}\left[1 + \dfrac{\cos \pi(|f| - f_1)}{2(B_T - f_1)}\right] & f_1 \leqslant |f| \leqslant 2B_T - f_1 \\[2ex] 0 & |f| > 2B_T - f_1 \end{cases}$$

The corresponding time-domain representation (that is, the impulse response) of this filter characteristic is given by:

$$p(t) = \mathrm{sinc}(2B_T t)\frac{\cos(2\pi\rho B_T t)}{1 - (4\rho B_T t)^2}$$

where $f$ is frequency,

$t$ is time,

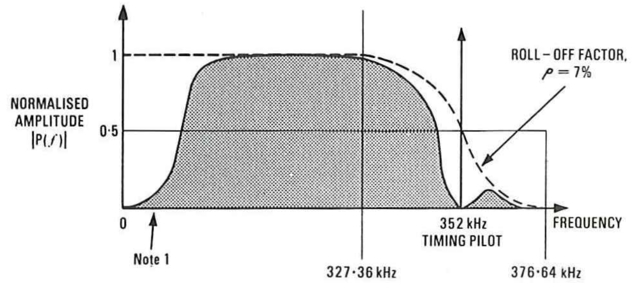
$B_T$ is the centre of the filter cut-off slope = 352 kHz,

$\rho$ is the roll-off factor = 0·07,

$f_1 = (1 - \rho)\,B_T = (1 - 0{\cdot}07) \times 352 = 327{\cdot}36$ kHz, and

$P(f)$ represents the Fourier transform of $p(t)$.

The resulting baseband spectrum, having been shaped by this filter and showing the 352 kHz timing pilot, is shown in Fig. 4.




*Note* 1: The low-frequency components have been reduced by the combined effects of scrambling the binary signal and then applying the technique of feedback balanced coding

FIG. 4—Schematic representation of baseband spectrum at the output of the send low-pass filter subsystem

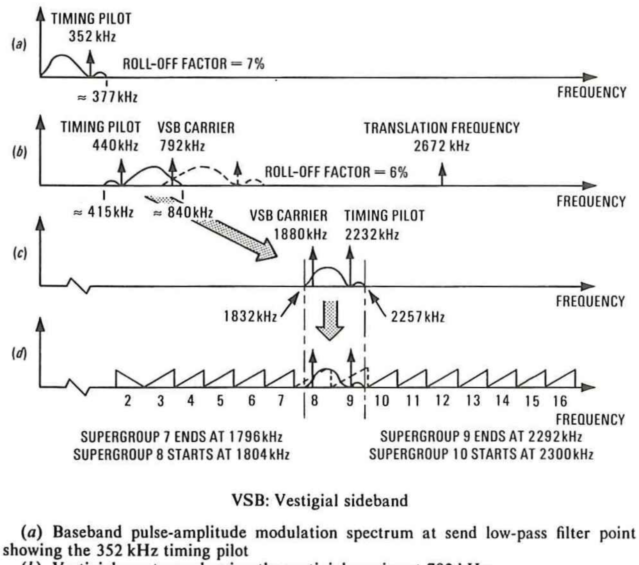
The design objective is that the raised-cosine filter should reduce the effects of intersymbol interference[3] to $\leq 1\%$ RMS in the absence of noise. This figure is based on practically realisable amplitude and delay distortion characteristics for the filter design.

### *Vestigial Sideband Modulator and Vestigial Sideband Send Filter Subsystems*

The vestigial sideband (VSB) modulator (VSBM) and VSB send filter (VSB SF) subsystems translate the shaped baseband signal (shown in Fig. 4) to an intermediate-frequency band by a VSB modulation technique. This process, which is shown in Fig. 5(*b*), has the following advantages:

(*a*) The VSB carrier (that is, 792 kHz) is present within the composite signal spectrum, and can, therefore, be extracted to provide coherent demodulation to baseband in the receiving modem.

(*b*) When the VSB signal is translated to the spectral location of supergroups 8 and 9 it is very easy to filter off the translation frequency and unwanted sideband thus produced. If, on the other hand, a direct translation from baseband to the spectral location of supergroups 8 and 9 had been carried out, the sidebands thus produced would appear very close to either side of the translation frequency;




VSB: Vestigial sideband

(*a*) Baseband pulse-amplitude modulation spectrum at send low-pass filter point showing the 352 kHz timing pilot
(*b*) Vestigial spectrum showing the vestigial carrier at 792 kHz
(*c*) Final translation to the spectral allocation of supergroups 8 and 9. The translation frequency (2672 kHz) and upper sideband are filtered off prior to the signal emanating from the analogue output port of the modem
(*d*) Typical hypergroup assembly of 15 supergroups. The modem signal can be seen to reside well within the bandwidth normally allocated to supergroups 8 and 9

FIG. 5—Spectral translations within the modem

this being a consequence of the low-frequency components of the baseband signal. Hence, it would then be extremely difficult to filter off the unwanted sideband and translation frequency.

### *Modulator and Oscillator Subsystems*

The oscillator (OSC) subsystem provides the 2672 kHz modulating frequency which is required to translate the VSB signal, via the modulator (MOD) subsystem, into the frequency band normally occupied by supergroups 8 and 9.

As shown in Fig. 2, the OSC subsystem is common to both the transmit and receive sections of the modem. It is therefore possible to have a slight frequency error between two modems. However, provided this error is kept within tolerable bounds, no significant impairment is incurred (since the critical frequency and phase sensitive process is during VSB demodulation, which, as stated earlier, is a coherent process).

### *Band-Pass Filter Send Subsystem*

The band-pass filter send (BPFS) subsystem does not require a very sharp band-pass characteristic to remove the 2672 kHz translation frequency and upper sideband produced by the MOD subsystem. Therefore, group-delay distortion effects are minimal.

The output from the BPFS subsystem is ready for insertion, via the ADD SG facility on the STE, into the FDM hypergroup. See Figs. 5(*c*) and 5(*d*).

## Analogue-to-Digital Signal Processing

Most of the analogue-to-digital signal processing is simply a reversal of the operations so far described for the digital-to-analogue direction; hence, the subsystems to which this applies are not discussed again. However, a few other subsystems are worthy of special note. From Fig. 2, these are:

### *Carrier Automatic Phase Control Subsystem*

The carrier automatic phase control (CAPC) subsystem is a phase-locked loop (PLL) circuit, which ensures that the VSB carrier frequency (792 kHz), after having been extracted from the demodulated supergroup 8 and 9 spectrum, maintains the correct phase relation for coherent demodulation to baseband.

### *Timing Automatic Phase Control (TAPC) Subsystem*

After the VSB spectrum has been demodulated to baseband, it is necessary to sample the 8-level analogue signal optimally in order to minimise the effects of intersymbol interference. To achieve this, the 352 kHz timing pilot is extracted to drive a PLL, which, in turn, produces the accurate timing signal required by the equaliser and the multilevel-to-binary converter (M–B CONV) subsystems.

The decoder subsystem also requires the extracted 352 kHz timing signal so that the framing symbol can be correctly identified. This is crucial in identifying the frames that have been subject to bit inversion during the application of the FBC operation in the transmitting modem.

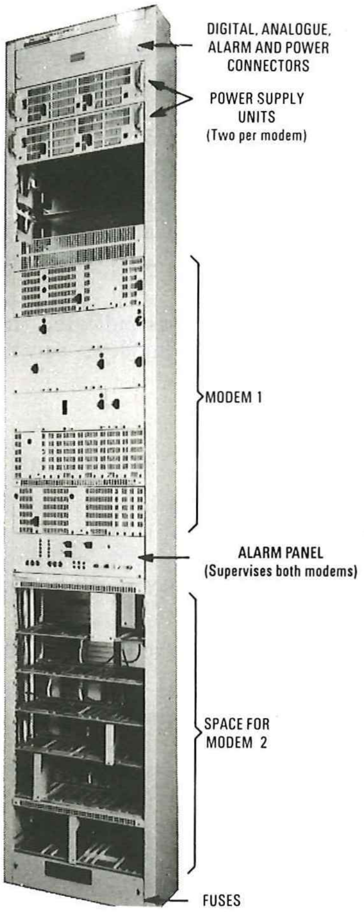
### *Equaliser Subsystem*

The modem uses a 41-tap transversal-type automatic equaliser. This equaliser attempts to track and compensate for any imperfections occurring in the incoming analogue signal; for example, attenuation and group-delay distortion, which are not only functions of frequency (primary dependence), but can also be time variant (say, because of seasonal temperature changes).

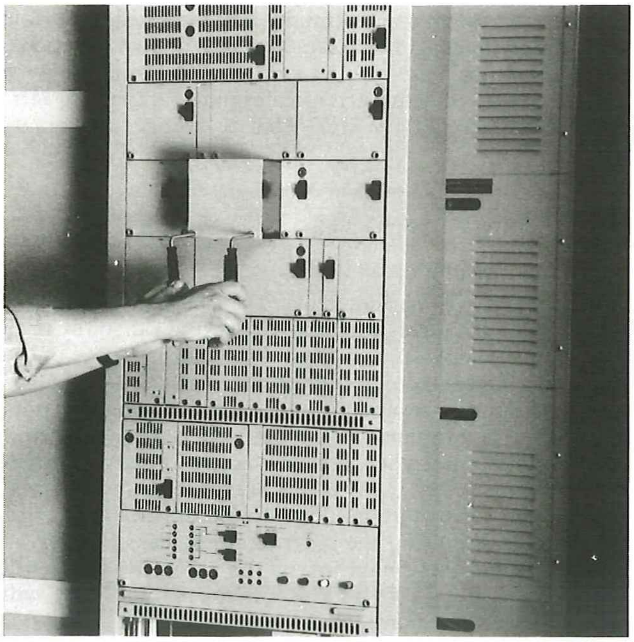
The equaliser functions in two modes, known as the *maximum level error method* and the *zero forcing method.*

These algorithms ensure the rapid convergence of the coefficients required for each of the taps of the equaliser in order to minimise intersymbol interference distortion[4, 5, 6].

For those FDM hypergroups which possess a very poor group-delay characteristic over the bandwidth of supergroups 8 and 9, it is possible to use a second group-delay equaliser. This fits directly into a pre-allocated slot in the rack which is normally blanked off. The second equaliser may be needed when the group-delay distortion over supergroups 8 and 9 exceeds 2·8 μs peak-to-peak.




(*a*) Modem rack equipped with one modem system



(*b*) Rack plug-in systems

FIG. 6—Modem rack

## OPERATIONAL AND INSTALLATION CONSIDERATIONS

### General Construction

The rack is fabricated from steel, and weighs approximately 100 kg unequipped. It can accomodate two complete modem systems, with one centrally located alarm unit monitoring both. Fig. 6(*a*) shows the rack equipped with one modem. Almost all of the subsystems of the modem (for example, COD, B–M CONV, etc.) are completely enclosed in a light-gauge steel casing passivated with zinc, and their plug-in position is clearly marked on each shelf of the rack. Fig. 6(*b*) shows one particular plug-in subsystem of the modem being extracted from the rack. The supervisory alarm subsystem can be clearly seen at the bottom of Fig. 6(*b*).

The rack wiring can be accessed only by removing rear panels, which prohibits back-to-back installations.

### Rack Dimensions

The rack is 2750 mm high, 520 mm wide and 250 mm deep. These dimensions are not exactly the same as current BT equipment practices; however, in the five sites where the rack has been installed (during the testing and evaluation phases), there have, as yet, been no significant accommodation problems.

### Power

The modem tolerates a supply voltage range of −42 V to −57 V (−48 V nominal), and consumes 300 W per fully equipped rack.

### Alarms

The modem indicates the local alarm conditions detailed in Table 1. Provision is made for these to be cabled to a remote location.

**TABLE 1**
**Alarm Conditions and Alarm Panel Indications**

| Alarm Indication | Conditions that may cause the alarm |
|---|---|
| REC S | Loss of digital input signal, or deterioration of the digital input signal to an HDB3 code violation count corresponding to a bit error ratio of $10^{-3}$ |
| SYNC | Loss of analogue signal, or severe degradation causing synchronisation problems |
| DIG ERR × 5 | Deterioration of the analogue signal that would result in a corresponding bit error ratio worse than $10^{-5}$ |
| DIG ERR × 3 | As DIG ERR × 5, except that this applies to a bit error ratio worse than $10^{-3}$ |
| SEND | An alarm signal that indicates the remote modem is in an alarm condition |
| FUSE and NVA (no volt alarm) | This alarm condition is generated in the case of (*a*) a blown fuse in one (or more) of the rack power supply units, or (*b*) a 'long-term' loss of power to the rack, causing the power supply units to latch in the OFF mode and, hence, requiring a manual reset |

## Maintenance and Fault Diagnosis

The kit supplied with each modem includes a set of test leads for local maintenance and fault diagnosis. These leads allow:

(*a*) access to the power supply units so that their output voltages can be checked against the quoted values, and

(*b*) access to a variety of signal monitor points.

The test leads provided for (*b*) are terminated in a special Japanese HF connector at one end, and a standard 75 Ω BNC connector at the other. These leads, in conjunction with an oscilloscope and/or spectrum analyser, can be used to observe various signal characteristics at the more critical stages of the signal processing found within the modem.

For example, a useful performance-indicating characteristic is the 8-level eye diagram produced from the recovered PAM signal; this can be examined both before and after equalisation at the receive low-pass filter (RLF) and multi-level-to-binary converter (M–B CONV) monitor points, respectively.

Since the eye diagram is an easily accessed signal (even when the system is carrying traffic), and contains a high degree of information about the error performance of the modem on a given hypergroup link, it is considered worthwhile to give a brief explanation of its most important features.

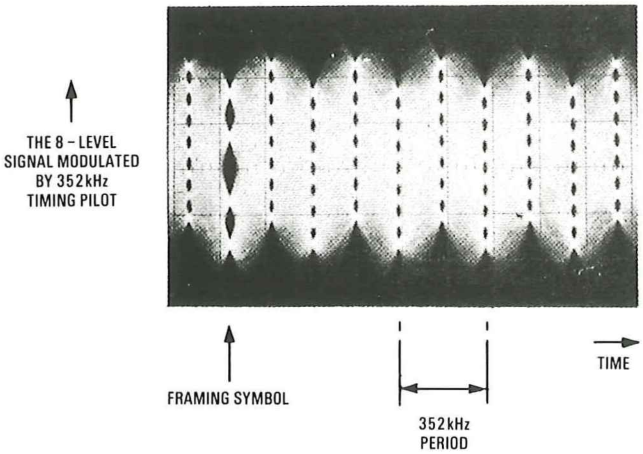
Fig. 7 shows a typical 'good-quality' 8-level eye diagram taken at the RLF monitor point.

The stacked eyes can be seen to have quite a small margin for error as a result of trying to maximise the information content of a given amplitude range and bandwidth; that is, power and spectral allocation constraints, respectively.

Errors in decision can arise from two primary effects:

(*a*) *Reduction of Eye Height Margin* Since each eye must share the assigned amplitude range, the allowable height of each eye is limited. Intersymbol interference and noise amplitude distortion effects must therefore be related to the allowable height of each eye; that is, the more levels that are used in a given amplitude range, the less is the eye height error margin.

(*b*) *Reduction of Eye Width Margin* The width of each eye (or timing error margin), is largely governed by the overall channel and filter frequency response. Restriction of the allowable bandwidth in the frequency domain leads to pulse spreading in the time domain. Hence, the 'tails' from adjacent signal sequences encroach further towards each other and thus restrict the tolerable timing error at the sampling points; for instance, this limits the jitter allowed on the recovered timing signal.




*Notes*: 1 The framing symbol can be seen constrained to reside at one of only four possible levels. This improves the error margin of the eye pattern for this symbol

2 The 352 kHz timing pilot can be seen as the the modulating envelope. This effectively raises the level of alternate sample points by one level

FIG. 7—Eight-level eye diagram observed at the receive low-pass filter monitor point

## EVALUATIONS

### Initial Laboratory Evaluation

Towards the end of December 1982, four modems were delivered to BT for evaluation.

A comprehensive test programme was carried out over the following weeks to see if the modem would be compatible with the normal networking requirements of BT. The analysis revealed two major problems:

(*a*) an unsatisfactory input jitter tolerance characteristic and too high a value for jitter gain; and

(*b*) an inability to generate an alarm indication signal (AIS) (that is, a binary all-ones pattern) when the digital input was removed.

The modem manufacturer has indicated, and in the case of (*a*) demonstrated, that the modems can be modified to resolve these problems.

### Field Evaluation

The modems have been tried over two routes:

(*a*) a microwave radio-relay link between Dalton (Cumbria) and Douglas (Isle of Man), and

(*b*) a coaxial cable link between Mondial international exchange (London) and Madley Earth Station (Herefordshire).

The field evaluation of each route consisted of:

(*a*) characterising the FDM link in terms of gain linearity, noise, carrier leaks, traffic density and types of signalling in use; and

(*b*) conducting a detailed error-monitoring programme over the rest of the test period.

The initial characterisation of the FDM link was necessary to allow the subsequently obtained error-performance results to be related to the particular link characteristics observed. Furthermore, by applying this procedure to all future field evaluations, it is hoped to achieve a greater understanding of the relationship between a given set of FDM link characteristics and an expected error performance.

Since the modem can essentially be considered to be data handling equipment, the characteristics of its operational error performance are very important. The parameter most commonly used to define error performance is the long-term mean error ratio (LTMER). This is defined, for binary signals, as:

$$\text{LTMER} = \frac{\text{number of binary errors observed in test period}}{\text{total number of bits in test period}}$$

However, the LTMER does not give any indication of how the errors are distributed over the measurement period. For example, the errors could have occurred randomly or in bursts. For the data user, a more detailed knowledge of the error statistics can be quite important.

This can easily be appreciated by considering the data throughput performance of a 're-transmission on error' system. In this case, where a whole block of binary information is re-transmitted if any errors are detected in the block, a bursty error distribution is preferable to a random error distribution comprising the same total number of errors.

To overcome this limitation of the LTMER, there will soon exist a revised CCITT Recommendation[7] that will specify the error performance in terms of:

(*a*) the percentage error-free seconds,

(*b*) the percentage of seconds where the error ratio $\geq 10^{-3}$,

(*c*) the percentage of 1-minute periods where the error ratio $\geq 10^{-6}$, and

(*d*) the percentage of time for which the system can be considered as unavailable (that is, periods of very high error ratio lasting several seconds).

It is intended that these parameters will provide more information to all network users about the error characteristics of the service offered. The measurement and subsequent presentation of the error characteristics obtained for the Madley–Mondial route, for example, were based on the parameters given above. However, because of a design limitation of the error-monitoring equipment used, the lowest error integration period possible was only 5 seconds (rather than the more suitable period of 1 second mentioned above). This reduction in the resolution of the errored integration period affects on the numerical value presented for these parameters; for example, the percentage of error-free 5-second integration periods must always be less than or equal to the percentage of error-free 1-second integration periods.

The results obtained for a 52-hour error-monitoring period are summarised in Table 2. For completeness, the LTMER measured over this period was approximately $8 \times 10^{-9}$.

**TABLE 2**
**Results of Error Monitoring for the Madley–Mondial Route**

| (*a*) Percentage of 5-second integration periods having | | | | |
|---|---|---|---|---|
| no errors | BER$\geq 10^{-7}$* | BER$\geq 10^{-6}$ | BER$\geq 10^{-5}$ | BER$\geq 10^{-4}$ |
| 99·8 | 0·2 | 0·142 | 0·018 | 0 |

| (*b*) Percentage of 10-minute integration periods having | | | | |
|---|---|---|---|---|
| no errors | BER$\geq 10^{-9}$* | BER$\geq 10^{-8}$ | BER$\geq 10^{-7}$ | BER$\geq 10^{-6}$ |
| 84·6 | 15·4 | 12·9 | 1·9 | 0 |

BER: Bit error ratio over given integration period

* Columns marked thus indicate the lowest BER resolution possible in the given integration period; that is, all integration periods containing $\geq$ 1 error are included in this column

*Note*: The results given in (*b*) were originally produced when it seemed likely that the CCITT would adopt a 10-minute integration period for the measurement of those periods over which the BER$\geq 10^{-6}$. However, as stated in the text, a 1-minute integration period has now been agreed internationally for this type of measurement, and it is not now possible to reprocess the results is this form

## THE EFFECT OF THE HYPERGROUP TRAFFIC LOADING AND POTENTIAL INTERFERENCE SOURCES ON THE ERROR PERFORMANCE OF THE MODEM

It is not sufficient merely to characterise the transmission quality of supergroups 8 and 9 in order to predict the expected error performance of the modem. The achievable error performance is also dependent upon other parameters such as the traffic loading on the hypergroup and external interference sources.

Before the more obvious potential sources of interference, such as impulsive noise, for example, are examined, it is interesting to note the effect of a normal hypergroup traffic loading on the error performance of the modem. This effect alone, as will be shown, makes the error performance of the modem quite difficult to predict accurately.

If a single voice-channel signal (from about 300 Hz to 3·4 kHz) is examined, a relatively large value for the peak-factor will be found; this means that large signal excursions, compared to the RMS value, are likely to occur quite frequently. However, if many such uncorrelated signals are combined, as is the case with a hypergroup (typically, 900 channels), then the overall signal will have a reduced peak-factor and the composite signal statistics will tend towards that associated with a normal distribution[8]. Thus, large signal excursions, compared to the RMS value, of this composite signal are now likely to be less frequent than the single-channel case.

However, there must still exist a finite probability that many channels will simultaneously exhibit individual large signal excursions from their respective RMS values. When this happens, the hypergroup amplifier may be temporarily overloaded and transient intermodulation products may be generated. If these intermodulation products fall into the bandwidth allocated to the modem, they may degrade the analogue signal sufficiently to cause the generation of digital errors.

Thus, the error performance of the modem can be seen to depend upon the statistical nature of the rest of the traffic on the hypergroup, and not just on the steady-state characteristics of the bearer bandwidth of supergroups 8 and 9. Consideration of this effect alone should make it apparent that the modem is unlikely to operate error free, in the long term, over a typical hypergroup assembly. In fact, the design specification for the modem indicates that an LTMER of $10^{-8}$ (or better) should be attainable for most hypergroup assemblies.

In addition to the effect described above, which may be classified as the *planned loading* effect of the FDM system, several other potential sources of interference could affect the analogue signal of the modem.

Some examples of these are:

(*a*) *Excessive FDM Loading* This, which can be considered as *unplanned loading*, can be caused by the levels of group and supergroup assemblies being incorrectly set, or by individual group or supergroup assemblies becoming noisy because of some fault condition. These conditions could cause excessive intermodulation products to be generated. This effect was observed during the Madley–Mondial field trial when two groups in supergroup 12 became extremely noisy for a period of 44 minutes. Intermodulation products were seen to fall in the bandwidth of supergroups 8 and 9 on a regular basis. The average bit error ratio exhibited by the modem for the duration of the fault was about $10^{-5}$.

(*b*) *Dry-Joints in an FDM Line System* These can cause transient breaks in the analogue FDM line signal. The duration of these breaks may be exacerbated by the modem losing alignment, and thus requiring a short period after the signal has been restored to settle back into normal operation. This is likely to be manifested as error-bursts probably lasting several milliseconds.

(*c*) *Impulsive Noise* This can be described as transient bursts of noise affecting the FDM system. Two potential sources of this impairment are lightning strikes and transient electromagnetic induction and radiation fields. The latter are usually associated with high voltage/current switching operations; for example, Strowger switchgear, relay-sets and the turning on or off of power supply units.

For true analogue-sourced signals carried on the FDM system, all of the effects mentioned above go virtually unnoticed unless particularly severe. For example, in the case of FDM speech the 'telephone activity factor[8]' ($T_L$) is only about 0·3, even in the busy hour; that is, for approximately 70% of the time a signal is not present and, hence, cannot be impaired (although noise bursts may still be heard). Furthermore, the redundancy present in the spoken word makes even severe impairment tolerable before intelligibility is lost. However, in the case of the analogue-encoded digital signal from the modem, channel occupancy is 100% (that is, $T_L = 1 \cdot 0$), and the redunancy of the signal is negligible in order to obtain efficient bandwidth usage. Hence, this signal is more likely to suffer irrevocable corruption, with the end product being digital errors.

## FUTURE WORK

Depending upon the availability of suitable FDM routes, it is hoped to carry out further field evaluations over the coming months. This, it is hoped, will give more operational experience and, hence, confidence of the likely error performance from different routes and different bearer media.

## CONCLUSIONS

While the transition from the analogue to digital trunk network takes place, there are likely to be occasions when it may prove useful to have the capability of routeing a 2048 kbit/s bit-sequence-independent digital signal over an analogue hypergroup assembly. These occasions are most likely to arise when:

(*a*) conventional digital plant does not exist and a digital transmission capability is required at very short notice (for example, the urgent provision of an X-stream service); or

(*b*) cost or technical reasons prohibit the replacement of an analogue transmission system before its planned obsolescence date(for example, this situation may occur with analogue satellite and submarine cable systems.)

The analogue signal from the modem lies within the bandwidth normally allocated to supergroups 8 and 9 of a hypergroup assembly. Access to the hypergroup assembly is obtained by the use of a wideband input/output facility, known as the ADD SG equipment, at the STE. Thus, only one modem can be used in a single hypergroup assembly. All STEs have the capability to accept the ADD SG equipment, but, because it is an optional facility, it is only likely to be found already fitted where requested when the STE was originally ordered.

From the limited field evaluations carried out so far, insufficient evidence has been gathered to make a definitive statement of the expected error performance of the modem over a given analogue link. Furthermore, this article has attempted to explain the difficulty in trying to make such a statement at all. That is, if it is assumed that the bearer bandwidth of supergroups 8 and 9 is of satisfactory quality, the operational error performance of the modem is largely governed by:

(*a*) The statistical nature of the rest of the hypergroup traffic; that is, the probability that this will cause the hypergroup amplifiers to be overloaded, and hence generate intermodulation products which affect the analogue signal of the modem.

(*b*) Transient noise producing mechanisms, such as relays (electromagnetic radiation) and dry-joints in the FDM network. Noise bursts produced by such mechanisms are not usually deleterious to FDM speech, but can irrevocably corrupt analogue-encoded digital data signals.

## ACKNOWLEDGEMENTS

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### Biography

Neil Harrison is an Executive Engineer in the Trunk Services Systems Engineering Division of BT National Networks. In Blackburn, in 1971, he began a 4-year apprenticeship with the local electricity supply authority. This was followed by a 6-year period of full-time study from which he initially obtained an OND in Electrical and Mechanical Engineering (from Blackburn Technical College) and, subsequently, a First Class Honours degree in Electrical and Electronic Engineering (from Preston Polytechnic). He joined BT in November 1981 and has, to date, been chiefly involved in advanced network performance studies.

## APPENDIX

## GRAY-TO-NATURAL-BINARY PRE-CODING

It is argued that the vast majority of noise corruptions of the received analogue signal (assuming a Gaussian distribution of noise), after demodulation and decoding back to a PAM format, usually result in only a single level change of the PAM signal.

If it is accepted that this postulate is valid, and it is further assumed that the original binary 3 bit symbols are already in a Gray code format, then a simple Gray-to-natural-binary conversion will result in only a single binary error from a single level decoding error of the PAM signal.

To illustrate the Gray-to-natural-binary pre-coding operation, and its implications, consider Table 3.

**TABLE 3**
**Pre-Coding Operation**

| Input Gray Code Symbol | Output Natural Binary Symbol | PAM Level |
|---|---|---|
| 100 | 111 | 7 |
| 101 | 110 | 6 |
| 111 | 101 | 5 |
| 110 | 100 | 4 |
| 010 | 011 | 3 |
| 011 | 010 | 2 |
| 001 | 001 | 1 |
| 000 | 000 | 0 |

Consider the 3 bit symbol 110. From an assumed Gray code format, this is converted to 100 in natural binary.

Thus, a PAM level of 4 would be sent forward for shaping and modulation prior to FDM transmission.

If noise, incurred during transmission, corrupts the decoded PAM signal to level 3 or 5 say, then these levels are decoded to 011 and 101, respectively (in natural binary). Now, when these are converted back into Gray code, symbols 010 and 111, respectively, are obtained. If these are now compared with the original 3 bit symbol, 110, it can clearly be seen that only a single binary error has been produced.

# Remote Monitoring of the Pressurised Cable Network

## Part 2—Cable Pressure Monitoring

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UDC 621.315.211.4

*This second part of the article[1] describes automatic monitoring of the pressures in telecommunication cables using small pressure transducers which can be inserted into cable joints. These transducers are connected to a single cable pair and the pressure information is monitored by a remote processor-controlled system.*

### INTRODUCTION

Most local, and trunk and junction cables terminating at telephone exchanges and repeater stations are pressurised with dry air to a maximum of 620 mbar (gauge) pressure*. (Dry air in this context means air with a dew point below −25°C.) This pressure serves to retard the ingress of moisture and enables maintenance staff, using suitable monitoring devices, to be alerted when a defect in a sheath occurs.

At the present time, the following devices are used to monitor the network:

(*a*) *Contactors* These are pressure switches fitted at intervals along the length of the cable; they are also fitted in equipment repeater cases.

(*b*) *Pressure Contact Gauges* These gauges are fitted on trunk and junction cables at the equipment cable pressurisation (ECP) rack and in the local network at the primary cross-connection points.

(*c*) *Flowmeters* These air-flow indicators are installed at the ECP rack.

### METHOD OF MONITORING CABLE PRESSURES

When a leak is present in a cable, the pressure inside it falls and this operates an alarm contact, which is either a contactor or the pressure contact gauge. This is received at the telephone exchange or the telephone repeater station where the internal maintenance staff test the circuit by a bridge technique to ascertain which contactor has operated. Fig. 1 shows the alarm circuit. The bridge technique fails if more than one contactor has operated.

When the testing is complete, the external plant maintenance control (EPMC) is informed by telephone. The external maintenance staff, in particular the precision testing officer (PTO), try to locate the leak in the problem cable.

The general distribution of the air pressure in a cable having a leak in the sheath is shown in Fig. 2. Fig. 2(*a*) depicts the distribution for a continuous-feed pressurisation system, while Fig. 2(*b*) is for a static system[2]. The leak point in both the continuous-feed system and the static system is shown by the discontinuity in the graph.

The leak is located by measuring the air pressure at intervals along the cable and drawing a graph of pressure versus distance. However, in a static system there is a problem, because the air pressure continues to fall while the measurements are being taken. These measurements create

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* 1 mbar = 100 Pa

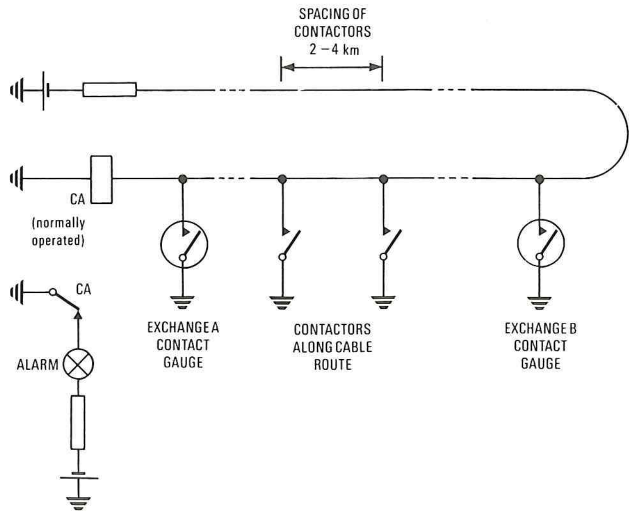
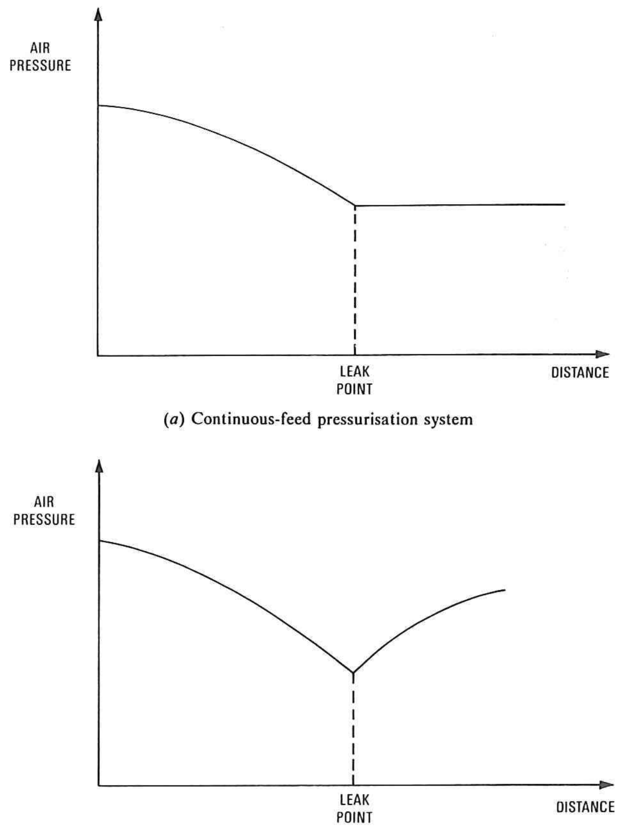


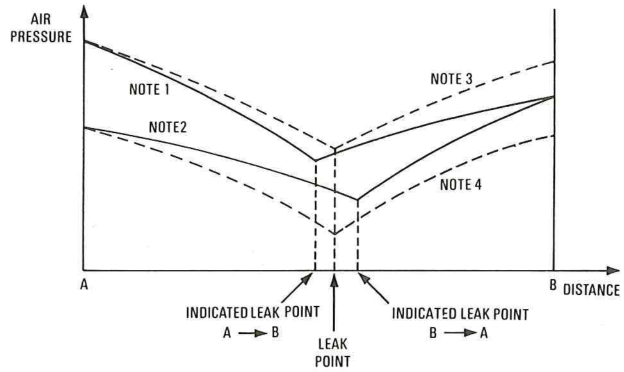

FIG. 1—Typical existing cable-pressure alarm system




(*a*) Continuous-feed pressurisation system

(*b*) Static pressurisation system

FIG. 2—Gas pressure distribution in a leaking cable




*Notes*: 1 Measured pressure gradient curve A → B
2 Measured pressure gradient curve B → A
3 Actual pressure gradient curve at begining of measurement period
4 Actual pressure gradient curve at end of measurement period

FIG. 3—Measured and actual pressure gradient curves for a leaking cable

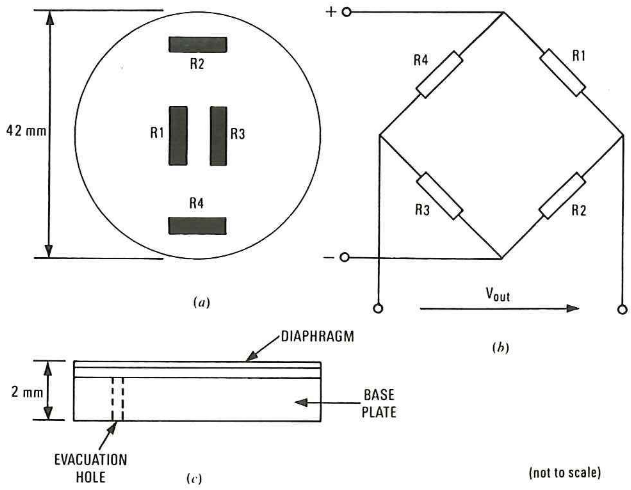



FIG. 4—Thick-film pressure sensor

an error in the minimum value of the graph, as this point may not correspond to the location of the leak. To minimise this error, a second series of measurements is taken at the same points as before except that this time they are taken in opposite sequence and another graph is drawn. If the two series of measurements occupy approximately the same time interval, then the mid point between the lowest pressure points of the two curves gives the approximate location of the leak, see Fig. 3. In practice, this is difficult since water may have to be pumped out of manholes, and difficulties may be experienced in parking the vehicle and setting up the safe road practices.

The problem of the two graphs can be solved if a series of simultaneous measurements can be taken. This is not practical with the techniques currently available to the PTO. What is required, therefore, is a pressure transducer installed at intervals along the cable so that each transducer can be addressed in sequence within a few seconds and the actual absolute-pressure information recorded by remote monitoring equipment at the telephone exchange. This obviates the time spent pressure testing on a double pressure run and tells the EPMC how quickly the cable is losing pressure. Bad leaks can be given priority while small leaks may be left for a day or two until the external maintenance staff are available to correct the sheath's defect.

## PRESSURE SENSORS

A sensor which senses the pressure change and which transduces this into a voltage or current is required to enable the physical quantity of pressure to be converted into an electrical quantity. This device is the pressure transducer. Two forms of pressure sensors which use the piezoresistive effect are now available. These utilise the technologies of thick-film and integrated circuits and, compared with other types of transducers, are small in size, have low power consumption (for example 10 mW) are compatible with the connection of electronic circuitry and, above all, are cheap as they are used extensively in the aircraft, process-control and automobile industries[3].

### Thick-Film Pressure Sensor

The sensing element consists of a circular edge-clamped ceramic diaphragm on which four thick-film resistors of the same geometry, connected in a Wheatstone-bridge configuration, are screened and fired (see Fig. 4(*a*))[4]. When the element is deformed by the difference in pressure between the two faces of the diaphragm, the resistors in the centre are elongated and so increase their resistance, while those at the edge are compressed and their value decreases. Fig. 4(*b*) shows the bridge connection that gives the maximum output voltage.

The diaphragm is bonded to an aluminium base-plate provided with a small hole (see Fig. 4(*c*)) that can be closed after the evacuation of the chamber between the two ceramic parts to obtain an absolute pressure transducer.

(Note: absolute pressures are required since, if the transducer is located inside the cable joint, the atmospheric pressure cannot be used as the reference; hence, the reference must be a vacuum and is related by

Absolute pressure = atmospheric pressure + gauge pressure)

### Integrated-Circuit Pressure Sensor

This type of sensor or transducer utilises the piezoresistive effect of silicon combined with integrated-circuit technology. Piezoresistivity is simply the change in resistivity of a material as a function of the applied stress. The silicon pressure transducer consists of a thin silicon diaphragm into which resistors are diffused (see Fig. 5)[3]. Mechanically, the resistors form part of the diaphragm, but electrically, they function independently, being isolated from the rest of the diaphragm by the p n junction. These resistors function as strain gauges similar to the thick-film transducer. This type of sensor is much smaller than the thick-film sensor.

Both types of transducer, but particularly the silicon type, require temperature compensation. Fig. 6 shows the basic gauge circuitry with a Wheatstone-bridge configuration. Compensation is required because of the various errors which arise from the temperature drift of the resistance-

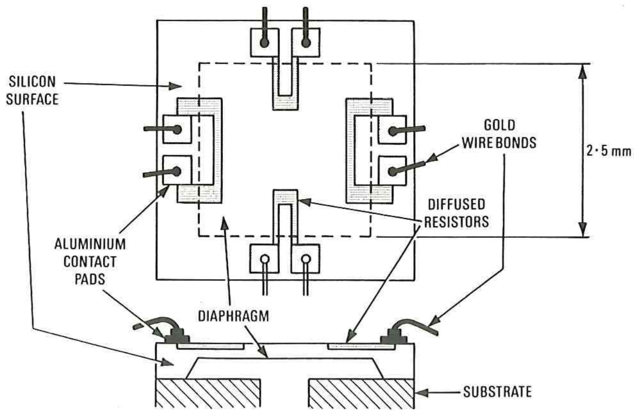



FIG. 5—Integrated-circuit pressure sensor

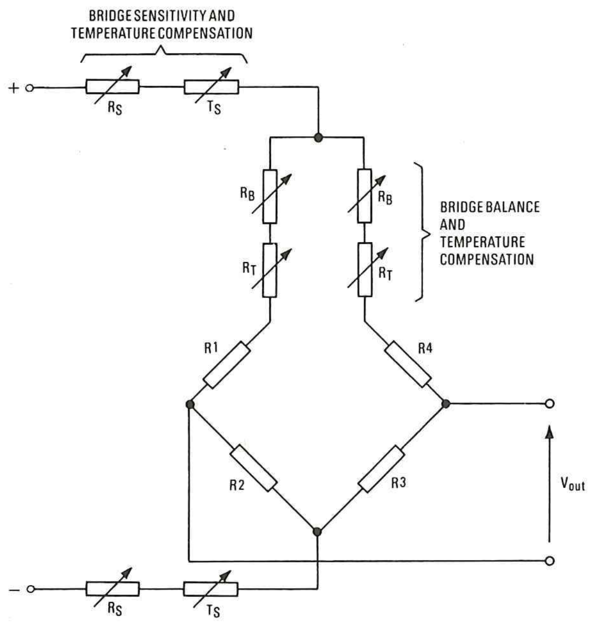



FIG. 6—Basic bridge circuit with compensating elements

element values. Compensating resistors and thermistors can be included in the sensor by the manufacturer. The external trim resistors can be precision elements or thick-film elements which are etched or trimmed. Manufacturers of the silicon type can add them to the chip by means of diffusion, ion-implantation or thin-film deposition techniques, and laser trim them for optimum performance.

As well as temperature compensation, both types of transducer require amplification, since the change in voltage output is of the order of 10–20 mV over the pressure range of interest. This amplification is achieved by using a differential amplifier which is buffered (often called an instrumentation amplifier).

## ADDRESSABLE PRESSURE TRANSDUCER

The current required to operate each transducer and its associated signal-conditioning circuitry is of the order of a few milliamps; thus, if all the transducers were continually powered, the voltage drop along the length of a cable pair would be excessive. To overcome this problem, and to enable each transducer to be individually monitored, an address recognition circuit (ARC) is required.

Fig. 7 shows the block diagram of the addressable pressure transducer (APT). The ARC, which uses complementary metal-oxide semiconductor (CMOS) components (that is, low-power integrated circuits), is powered continuously from the line, awaiting a train of address pulses. If the address received compares with the preset address of the APT, a power-up signal is applied for approximately 2 s to the pressure sensor circuit.

The APT is connected to both the A- and B-wires of the monitoring pair via a diode bridge, in order for the APT to be non-polarity conscious. Any fluctuations of voltage on the line caused by the address pulses are sensed by the CMOS logic circuitry via the line-tap resistor R1 and capacitor C1. These address pulses, which are applied to the line at the remote monitoring unit (RMU), are shifted into a register whose output is connected to a logic comparator. This comparator holds the identity or address of the APT. On recognition of the correct address, the 2 s power-up pulse is connected to the pressure-sensor circuit, which powers up from the line via resistor R2.

The pressure is sensed by the transducer, amplified and connected to a voltage-to-frequency converter. The output

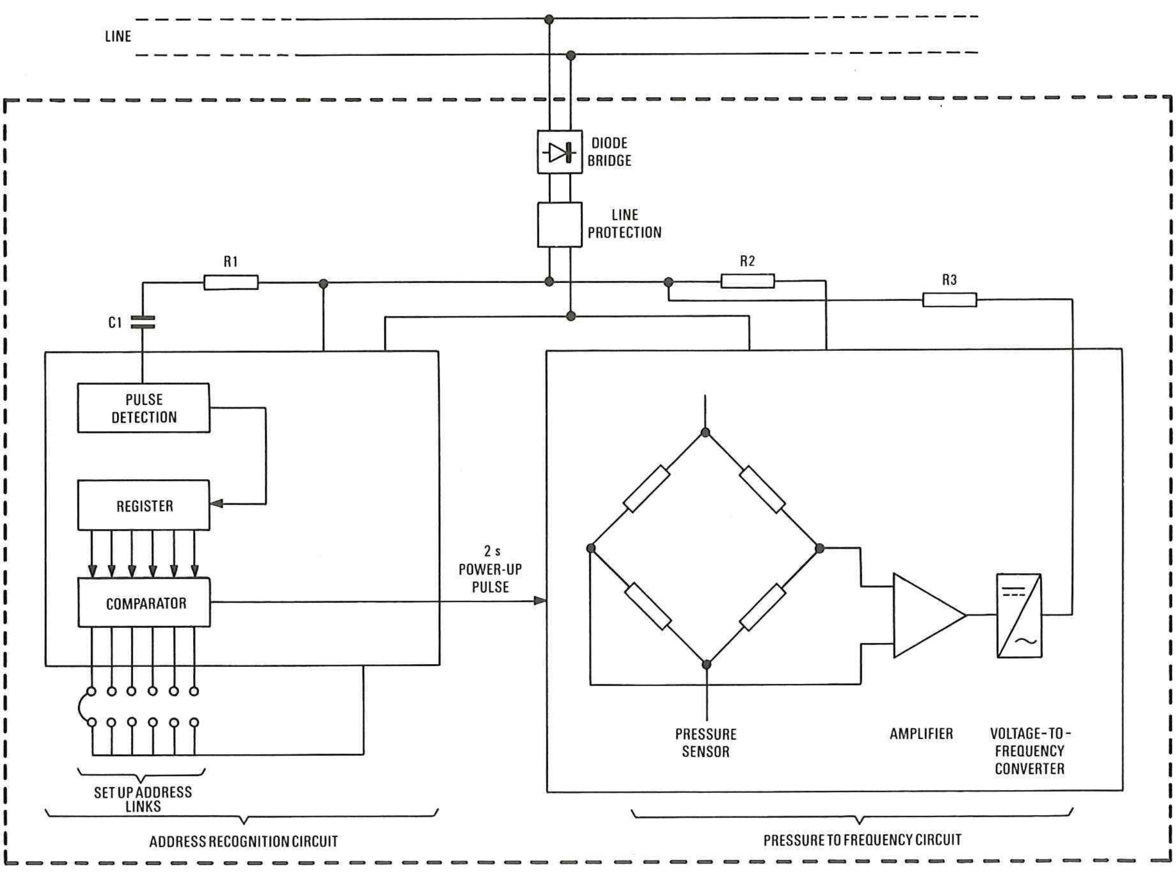



FIG. 7—Block diagram of addressable pressure transducer

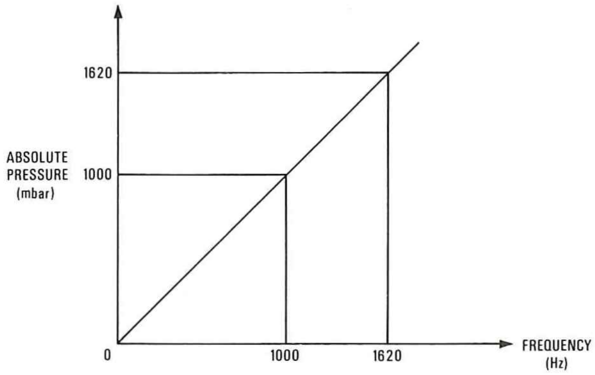



FIG. 8—Pressure-to-frequency conversion graph

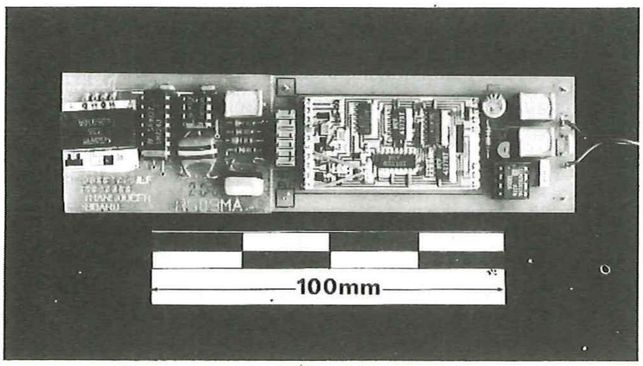



FIG. 9—Addressable pressure transducer

signal modulates the line voltage via resistor R3 at a frequency directly proportional to absolute pressure. (See Fig. 8). This signal is received by the RMU at the telephone exchange.

The APT shown in Fig. 9 is small enough to fit inside a cable joint; thus, no pipe connections or fitments are required.

## REMOTE MONITORING UNIT

The RMU (see Fig. 10) comprises line interfacing circuits and a microprocessor to control the output of the address

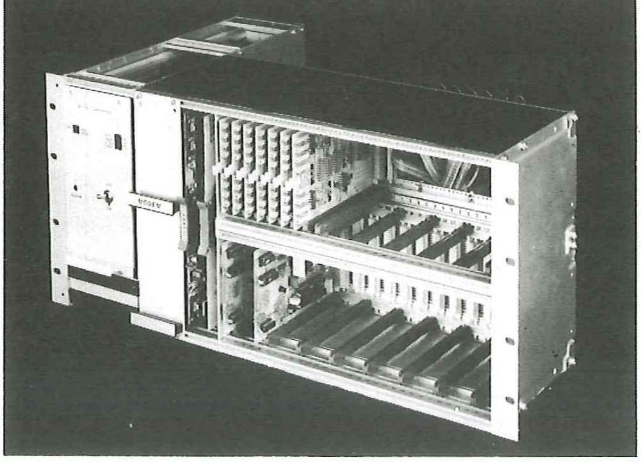


FIG. 10—Remote monitoring unit

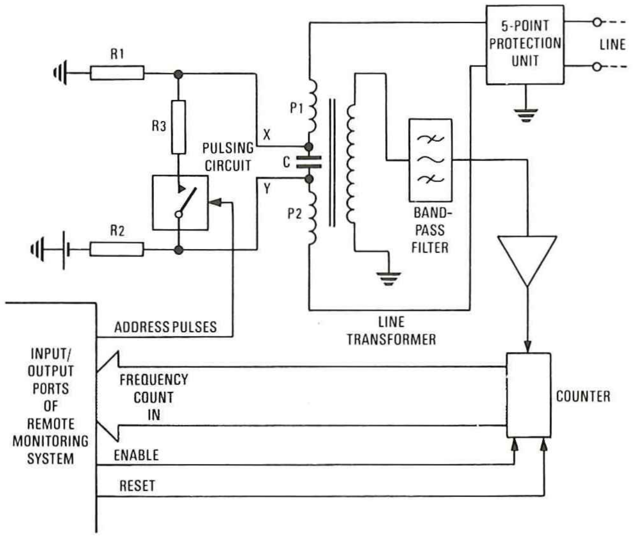



FIG. 11—Line interfacing circuit

pulses and to receive the line signals. Fig. 11 shows a simplified diagram of the RMU line interface circuits.

The exchange battery (−50 V) and earth (0 V) are connected via the resistors, R1 and R2, to the primary windings, P1 and P2, of the line transformer via the pulsing circuit. Capacitor C acts as a DC block.

The output of the primary windings is connected to the A- and B-legs of the monitoring pair via a 5-point protection module situated at the main distribution frame in the telephone exchange.

On power-up, the addressed APT sends a signal at a frequency proportional to the absolute pressure. At the RMU, this signal is transformed into the secondary of the line transformer, then filtered, amplified and gated into the counter for exactly one second. This count is, therefore, a time window having a value that equals the frequency in hertz. The count is down loaded when required into the processor memory via the input ports for future processing.

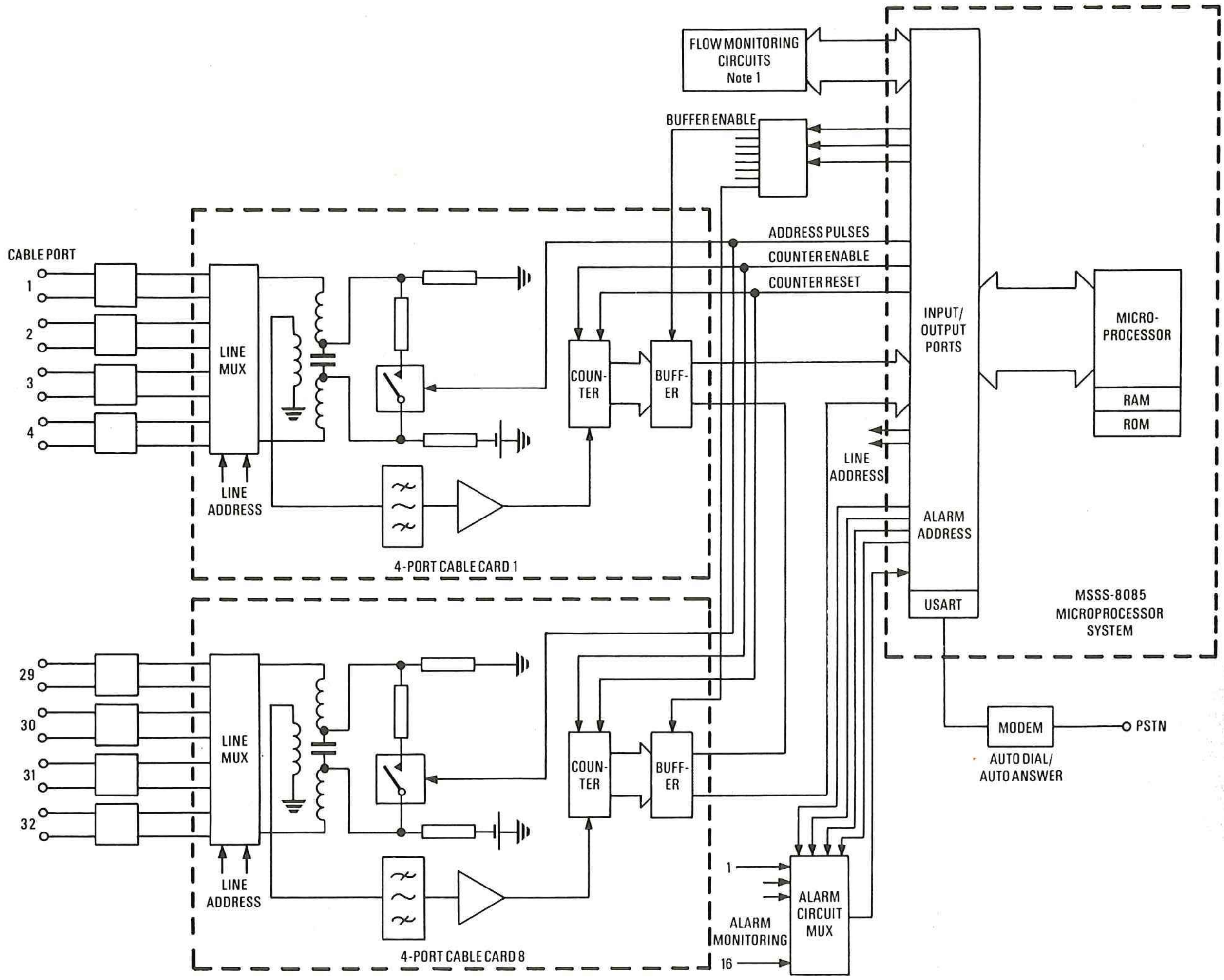
All the APTs are continually monitored. Whenever pressure falls below a preset value, information is sent to the EPMC via the dial-up modems. The EPMC officer can also interrogate the RMU to output any cable-pressure readings of interest.

A total of 32 cable ports with a maximum of 32 APTs per cable port allows 1024 transducers to be monitored. More than one cable pair may be connected to a cable port. The maximum distance that an APT can be installed from the RMU is governed by a maximum loop resistance of 2000 Ω. As well as addressable pressure transducers, addressable humidity contactors can be installed on the same cable pair if required. Humidity sensors are installed in many repeater housings.

Fig. 12 shows the block diagram of the complete RMU. The monitoring of the cable flows were described in Part 1 of this article[1]. Included in the system is the facility to monitor up to 16 alarm circuits, either 'contact' or 'earth' alarms. This is useful for monitoring the low-pressure and humidity alarms of the compressor/dessicator units installed at the telephone exchange or repeater station.

## CONCLUSIONS

A cable pressurisation monitoring system has been developed that automatically monitors the air flows and pressures of the cable network and reports to the EPMC when an alarm condition exists. When the EPMC is unmanned, these alarms are diverted to the 24-hour repair service control (RSC).




*Note* 1: See Part 1 of this article

MUX: Multiplexer
PSTN: Public switched telephone network
RAM: Random-access memory
ROM: Read-only memory
USART: Universal synchronous asynchronous receiver transmitter

FIG. 12—Block diagram of remote monitoring unit

The engineer in the EPMC or the RSC can interrogate any RMU to obtain information on the current pressurisation state of any cable.

The piezoresistive pressure transducer has proved to be a reliable device in the field. Its small size, compatibility with electronic circuitry and its low cost has led to the development of an addressable pressure transducer which is small enough to be located inside a cable joint. Pressure readings are continually monitored and can be received at the EPMC in order to observe the simultaneous readout of the pressures along the cable. These enable the EPMC officer to determine the possible location of a leak, and to send the engineers to the problem cables requiring the highest priority; that is, cables which are losing pressure fastest. Thus, by remotely monitoring the pressurised cable network, the increase in practical information allows the external-cable engineers to manage the network with a greater degree of control and efficiency.

## ACKNOWLEDGEMENTS

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## Biography

Jim Smith joined the Post Office in 1966 as a Trainee Technician (Apprentice). After a spell of maintenance duties in Dundee Area, he won a minor/major award to study at Heriot-Watt University where he gained his degree in Electrical and Electronic Engineering. Regraded Executive Engineer in 1978, he joined the External Plant Research Section of BTRL. He is a member of the IEE.

# Factors Involved in Determining the Performance of Digital Satellite Links

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*This article summarises the techniques available for implementing digital satellite transmission systems, and discusses current and probable new error-performance objectives that will be consistent with the requirements of the integrated services digital network. It also discusses factors affecting performance, including capacity/quality trade-offs, satellite and earth station parameters, atmospheric propagation conditions, interference and intermodulation. It concludes by indicating briefly the techniques available to a system designer for meeting performance requirements.*

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## INTRODUCTION

A great deal of work has been carried out recently on the definition of the integrated services digital network (ISDN). Some of it has been directed at laying down transmission-performance standards, and a main task has been concerned with identifying error-performance objectives for an end-to-end connection. It has been necessary, as part of this work, to consider the role in such a connection of all possible transmission media, including satellites, and the result has been the determination of error-performance requirements for such media. This is a new approach to the derivation of performance objectives, especially for satellite systems where current guidelines were derived from previous analogue requirements[1] without direct guidance from the CCITT‡.

This article briefly discusses the use of digital techniques on satellite systems, and compares existing error-performance requirements with the new ISDN requirements. The main body of the article describes the factors affecting the performance of satellite systems, and the conclusion briefly lists the techniques available to system designers for accommodating these factors.

## USE OF DIGITAL TECHNIQUES IN SATELLITE SYSTEMS

At present, most of the traffic on satellite systems is carried by using analogue techniques. Usually, frequency modulation (FM) is used, and multiple access to a single repeater (transponder) is achieved by frequency-division (FDMA). However, this is now beginning to change and by the end of the decade, a large proportion of satellite traffic (probably around 50%) will be carried by using digital techniques. The main reason for this increase in digital capacity is the introduction of time-division multiple-access (TDMA) by INTELSAT (the global International Telecommunications Satellite organisation) and by EUTELSAT (the European Telecommunications Satellite organisation). Both organisations are scheduled to commence TDMA operation in 1985.

### Time-Division Multiple-Access

TDMA is an alternative to FDMA for giving a number of users access to a single transponder. In essence, it operates by each participating earth station providing a buffer so that relatively low-rate continuous digital information coming into the earth station from the national telephone network can be transmitted to the satellite in bursts at a very high bit rate. The periodicity at which these bursts are transmitted is the *system frame rate.* As an example, an earth station operating in a TDMA system with a 2 ms frame period and 120 Mbit/s transmission rate may be fed with digital information at 10 Mbit/s. (This could be, for instance, about one hundred and sixty 64 kbit/s voice channels.) At 10 Mbit/s, the station receives 20 kbit per 2 ms, which, at a transmission rate of 120 Mbit/s, requires a burst of 167 $\mu$s duration. The remainder of the 2 ms frame is then free for other users of the transponder to transmit their bursts. An essential feature of TDMA is, of course, the need to ensure that only one station transmits at a time. This is achieved by using a synchronisation system which is based on the use of reference bursts. These are short bursts, transmitted once per frame by specially designated reference stations, which indicate the start of the frame. Each earth station is allocated a frame position relative to the reference burst and, by observing the actual position of its burst relative to the reference burst by some form of loop-back arrangement, is able to maintain correct timing. These broad principles of TDMA are illustrated in Fig. 1. Further discussion of the complexities of TDMA is beyond the scope of this article[2, 3, 4].

### Other Digital Techniques

Although it will be a major system, TDMA is not the only means of providing digital transmission capacity. The use of FDMA, as described above for FM carriers, is just as readily applicable for digitally modulated carriers, and systems using this technique can implement from a few to thousands of channels on each carrier. Where just one channel is provided on each carrier, the system is referred to as *single-channel-per-carrier* (SCPC), and this technique is popular for low-capacity systems such as those implemented on some domestic and business satellites[5]. INTELSAT has operated an SCPC system for many years using phase-shift keying (PSK) modulated carriers. A variation of SCPC is single-channel-per-carrier PCM multiple-access demand-assigned equipment (SPADE)[6]. This title summarises the essential features of the equipment very well, even if the acronym is somewhat contrived. Demand assignment indicates that capacity on the system is allocated to a particular earth station only when it is required.

The opposite end of the scale to SCPC is a carrier with so many channels that it occupies the entire available

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* LEWIS, J. R. Factors involved in determining the performance of digital satellite links. *The Radio and Electronic Eng.*, Apr. 1984, **54**(4), pp. 193–198.

‡ CCITT—International Telegraph and Telephone Consultative Committee

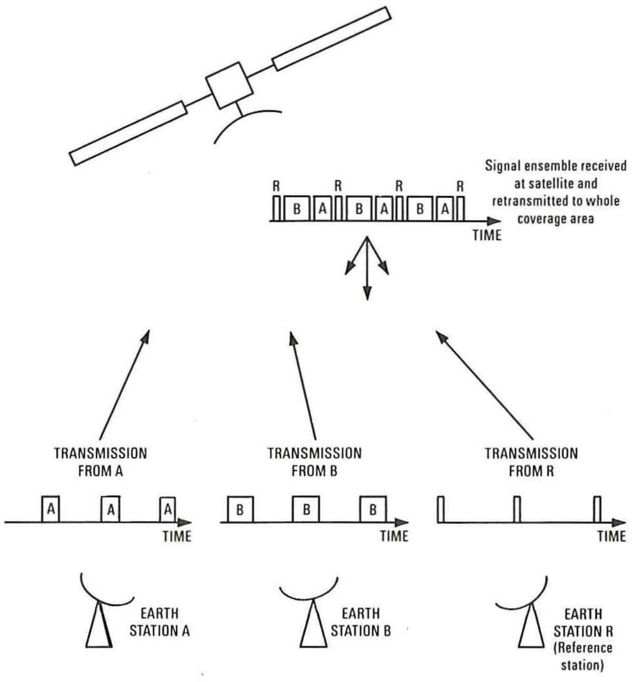



FIG. 1—Principles of TDMA

bandwidth of the transponder. In this case, neither FDMA nor TDMA is implemented, since just one station uses all of the bandwidth all of the time. Digital carriers of this type can provide very high capacities, and are likely to find application for carrying very large traffic streams between two points; that is, the satellite transponder effectively becomes a cable in the sky. Their first application is likely to be between Europe and North America towards the end of this decade.

A further arrangement that can be mentioned is combined TDMA and FDMA. In such a system, a transponder accommodates several carriers of, say, 2 Mbit/s, on an FDMA basis, and each carrier then provides capacity for a number of smaller users (perhaps 64 kbit/s each) on a TDMA basis. This sort of arrangement is likely to find application in business satellite systems.

From the above, it can be seen that a wide variety of transmission techniques is available for implementing digital transmission systems via satellites. The choice of the type of system to be used requires consideration of many factors including traffic levels, number and location of participants, cost, services to be carried, etc. The availability of such a range of techniques contributes greatly to the flexibility of satellites as a transmission medium.

## PERFORMANCE REQUIREMENTS FOR DIGITAL SATELLITE SYSTEMS

### Current Error-Performance Requirements

The most important transmission parameter for any digital transmission system is the bit error rate (BER), and requirements for this and many other satellite system parameters are the subject of Recommendations of the CCIR†. The current objectives are contained in Recommendation 522[7], the essential features of which are summarised in Table 1. Recommendation 522 has been in existence for some time now, and was designed to provide suitable transmission performance for digitally-encoded voice telephony traffic.

† CCIR—International Radio Consultative Committee

**TABLE 1**
**Summary of the Main Requirements of CCIR Recommendation 522**

**Allowable bit error rates at the output of the hypothetical reference digital path for systems in the fixed-satellite service using pulse-code modulation for telephony**

| Bit error rate | Integrating period | To be achieved for |
|---|---|---|
| $10^{-6}$ | 10 minutes | at least 80% of any month |
| $10^{-4}$ | 1 minute | at least 99·7% of any month |
| $10^{-3}$ | 1 second | at least 99·99% of any year |

### ISDN Error-Performance Requirements

Since Recommendation 522 was written, a significant amount of work has been done on defining the ISDN. One aspect of the work has been agreement on end-to-end error performance of ISDN connections, the results of which are laid down in CCITT Recommendation G821 (see also reference 8). This Recommendation, which has been revised extensively recently, now includes indications of requirements for satellite systems used to implement part of an ISDN connection. As a result of the need to accommodate a range of services in addition to voice telephony, as well as the wish to provide the best practicable transmission quality on the ISDN, these satellite requirements are rather more demanding than those of Recommendation 522. It is clear that the CCIR will need to consider these new requirements, and it seems likely that new objectives will be substantially agreed during the 1982–86 study period. Indeed, the first draft of text for a new Recommendation was agreed at the April 1984 Interim Meeting of CCIR Study Group 4. This text differs from Recommendation 522 in many detailed aspects, but one main difference is the adoption of a long-term BER requirement of $10^{-7}$ for 95% of the time as one option. There is, in fact, no practical limit to the performance that could be provided by using satellite transmission, but since improved performance is achieved only at the expense of reduced capacity and, hence, increased cost (as discussed in more detail below), the requirements will be set only as high as they really have to be.

### Other Error-Performance Requirements

One other set of BER requirements is those applied to several business satellite systems. Because these systems do not need to provide the high capacities required of main-service international systems, it is possible to take advantage of a trade-off between the quality and capacity of the system. The result has been the general adoption of a design criterion of BER = $10^{-6}$ for 99% of the time, although this is not the subject of any CCIR Recommendation.

### Clock Stability

Along with error performance, the other crucial digital system parameter is clock stability. For any satellite system other than TDMA, this parameter does not have much impact, because the network clocks used to feed data to the earth station are also used to transfer the data across the satellite link, thus making the link effectively transparent to the clock. The only consequence of the satellite link is that small regular movements of the satellite relative to each earth station result in Doppler-effect variations in the carrier frequency, causing a clock frequency wander to be imparted to the signal at the receiving earth station. However, this wander is cyclic, integrating to zero over a 24-hour period, and can, therefore, be removed by adequate buffering; it does not lead to timing slip.

The situation is different with TDMA systems. The rate at which information is transferred on a TDMA system is governed by the clock at the reference station, and this is usually independent of the clocks controlling data in and out of any two earth stations communicating over the system. This results in a need for interfacing TDMA and national network clocks at each earth station, with the inevitable consequence of slip occurring. However, most (if not all) TDMA systems will be controlled by clocks meeting the stability requirements laid down in CCITT Recommendation G811 (that is, 1 part in $10^{11}$) and slip will, therefore, be restricted to the value given in that Recommendation[9].

## FACTORS AFFECTING SATELLITE SYSTEM PERFORMANCE

This section discusses the various factors that have a bearing on the error performance of a satellite system and demonstrates how, in most cases, they result in a quality/capacity trade-off.

### Quality Against Capacity

The most striking way to demonstrate this trade-off is by considering a basic design equation of a digital system:

$$C/N_0 = (E_b/N_0) \times b \qquad .....(1)$$

where

$C$ = carrier power
$N_o$ = noise power per hertz
$E_b$ = energy per bit
$b$ = bit rate.

The ratio $C/N_0$ is a fundamental parameter of any satellite system. It depends on satellite and earth station parameters (for example, transmitter power, receiver noise and sensitivity) and on the environment in which the system operates (for example, propagation conditions and interference). Given a satellite, earth stations and defined operating conditions, the $C/N_0$ tends to be a constant characterising that particular system.

The ratio $E_b/N_0$ is the well-known measure of quality of a modulated digital carrier at the input to a demodulator and, for any particular modulation method, it is possible to draw curves of $E_b/N_0$ against BER on the demodulated digital sequence (see Fig. 2). It is therefore possible to regard $E_b/N_0$ as a direct function of transmission quality.

The final element of the equation is the bit rate, $b$, which is clearly a direct measure of system capacity.

This equation shows, therefore, that for a given satellite system characterised by a particular value of $C/N_0$, the systems designer is free to trade quality for capacity by adjustment of $E_b/N_0$ and $b$. (It is worth pointing out that this trade-off is also impacted by other factors, such as the limitation of bit rate by available bandwidth and the fact that the $E_b/N_0$ versus BER curve is, in fact, dependent on the bit rate at which the demodulator operates. However, these and other similar effects can be disregarded for the purposes of the point being illustrated.)

This example illustrates the flexibility available to a system designer for a fixed $C/N_0$. There is also the need to determine a suitable $C/N_0$ value in the first place, given that approximate required values of $E_b/N_0$ and bit rate (that is, quality and capacity) will be known, and the following paragraphs highlight the major system parameters that have an impact on this figure.

### Link Design Analysis

The $C/N_0$ figure used above is the ratio of carrier power to total noise power per hertz at the input to the demodulator. The carrier power is a function of both up-link (that is, earth-to-satellite) and down-link parameters, and $N_0$ is the sum of all noise (thermal, interference and intermodulation) on both up- and down-links. The ratio is therefore best considered in its constituent parts:

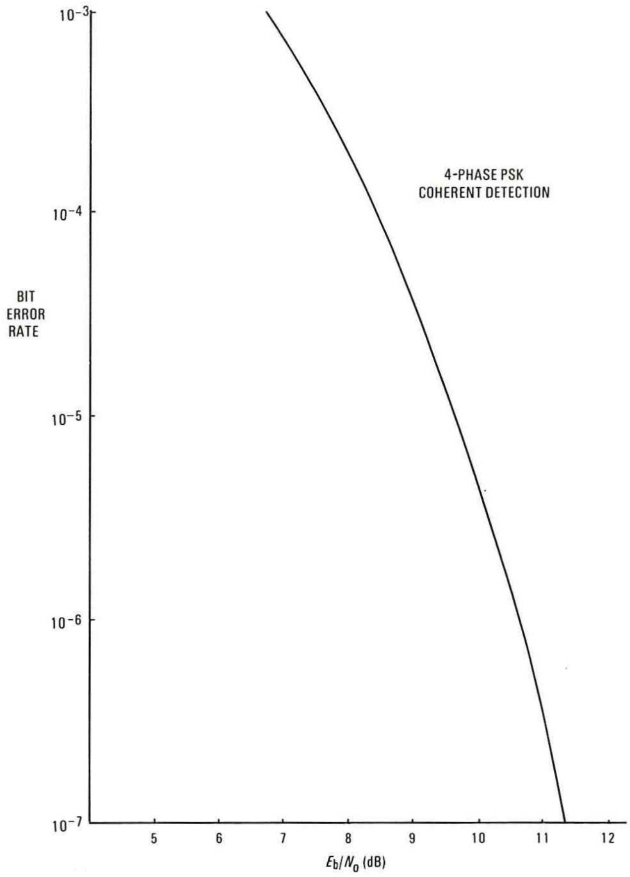



FIG. 2—Graph of demodulated BER against $E_b/N_0$ at demodulator input (Theoretical)

$$C/N_0 = -10\log\{(C/N_{tuo})^{-1} + (C/N_{tdo})^{-1} + (C/N_{io})^{-1} + (C/I_{uo})^{-1} + (C/I_{do})^{-1}\} - M \text{ dB} \qquad .....(2)$$

where

$N_{tuo}$ = up-link thermal noise per hertz
$N_{tdo}$ = down-link thermal noise per hertz
$N_{io}$ = equivalent intermodulation noise per hertz
$I_{uo}$ = equivalent up-link interference noise per hertz
$I_{do}$ = equivalent down-link interference noise per hertz
$M$ = margin to allow for degradation due to high-power amplifier non-linearities, expressed in decibels.

It should be pointed out that in equation (2), as in all satellite link design work, the various $C/N$ parameters would normally be expressed in decibels and would thereby require an antilog process as well as addition.

A further point worth noting is the reference of all noise contributions to a common carrier power $C$. This is only for mathematical convenience; in reality up- and down-link power levels are very different, typically by between 30 and 50 dB, depending on the system.

### Satellite and Earth Station Parameters

If each of these elements is taken in turn, it is possible to consider $C/N_{tuo}$ and $C/N_{tdo}$ together, since they can both be described by the following expression:

$$C/N_{tuo} \text{ (or } C/N_{tdo}) = \text{EIRP} - A - L + G/T - k \qquad .....(3)$$

where

EIRP = equivalent isotropically radiated power, that is the power transmitted in the direction of the satellite (earth) from the earth station (satellite). It is a function of transmitter power and antenna gain relative to isotropic.

$A$ = atmospheric attenuation.

$L$ = path loss between isotropic antennas. It is a combination of spreading loss, which reduces transmitted power to the power per square metre at the surface of a sphere with a radius equal to the satellite-to-earth station distance, and the theoretical isotropic aperture value for the receive antenna at the operating frequency.

$G/T$ = figure of merit for the satellite (earth station) receive system. $G$ is the receive antenna gain relative to isotropic and $T$ is the receive system noise temperature referred to the input of the low noise receiver. $T$ is much higher for the satellite, looking at the 'warm' earth, than for the earth station looking at 'cold' space. However, rain and clouds increase the earth station $T$, making the earth station $G/T$ weather dependent.

$k$ = Boltzman's constant.

If the parameters $T$ and $k$ were to be removed from equation (3), the resulting equation, when applied to the up-link, would give a power $C$ at the satellite input. The value of $C$ necessary to drive the satellite power amplifier, usually a travelling wave tube (TWT), to deliver maximum (saturated) output power can be called $C_s$ and the value of $C_s$ for any particular satellite is a measure of its sensitivity. Satellite transmitters are frequently operated below saturated output to reduce intermodulation effects, and the difference between $C_s$ and the $C$ value used in practice is known as *input back-off*, the resulting difference between actual and maximum transmitted power being known as *output back-off*. Fig. 3 illustrates the relationship between input and output back-off for a typical satellite TWT. The amount of back-off used in any system is referred to as the *satellite operating point*, and it can be seen from equation (3) that this can be controlled by appropriate choice of earth station EIRP. However, it can also be deduced from equation (3) that as the satellite is made more sensitive (that is, as $C_s$ is decreased), the EIRP required reduces and hence $C/N_{tuo}$ decreases.

## Path Loss

From the above discussion, some idea can be gained of the way in which main system parameters such as transmitter powers, receiver noise temperatures and sensitivities impact on system performance and capacity. However, the effect of such factors is almost intuitively obvious; what is more interesting is to look at the impact of $A$ and $L$ in equation (3). Taking path loss $L$ first, it has already been pointed out that this includes the isotropic aperture, a factor that is frequency dependent. The result is that path loss increases with frequency, the difference being about 7 dB between the two currently used up-link frequency bands at 6 and 14 GHz. This degradation with increasing frequency can be considered to be compensated by transmit and receive antenna gains which increase with frequency, but this is for fixed antenna diameter and is accompanied by a reduction in beamwidth (and hence coverage area). This may or may not be acceptable, depending on application. It is not possible, therefore, to generalise on the net effect of higher frequency operation on clear-weather system performance, although it can be said that it usually results in the use of smaller antennas, both on the satellite and on the ground. The fact that path loss also varies with the distance from the satellite to earth station has only a small impact, with only a few tenths of a decibel variation in $L$ occurring across the service area of the satellite system.

## Atmospheric Attenuation

The other factor of interest in equation (3) is the atmospheric attenuation $A$. This is a difficult factor to allow for in system design because it varies not only with time, but also from one climate to another and is, therefore, much more of a problem in some countries than in others. It is also frequency dependent, being more severe at the 14/11 GHz band than at 6/4 GHz. During clear-sky conditions, which apply for 80–90% of the time in most climates, attenuation is very low, probably below 0·5 dB even at 14 GHz. However, during cloudy conditions, it starts to rise and, during precipitation (rain, snow etc.), it can become very high. The effect is at its worst during very heavy rain storms, with values of 10 dB frequently being measured for short periods in the UK (higher values have been recorded) and in wetter climates values near 20 dB have been measured at 14 GHz. Clearly, it is not practicable to design systems to give normal performance during short periods of such heavy fading, and the problem is to determine the percentage of time during which degraded performance is tolerable. It is this sort of consideration that results in the form of the performance requirements quoted in Table 1.

In order to determine appropriate values of attenuation to use in system design, long-term statistics, which show attenuation exceeded against time percentage, are used. An example of such a curve is given in Fig. 4. These statistics also vary from month to month, but for any climate, it is possible to identify the worst month, and the models used

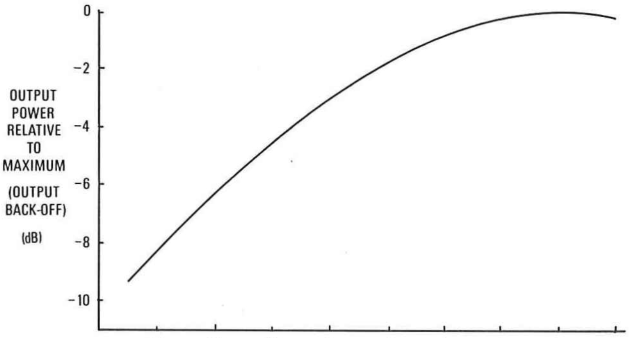
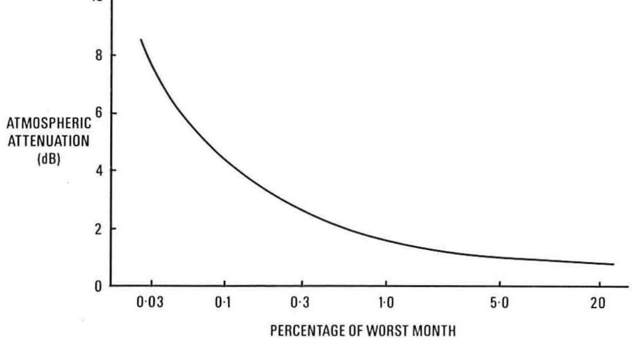
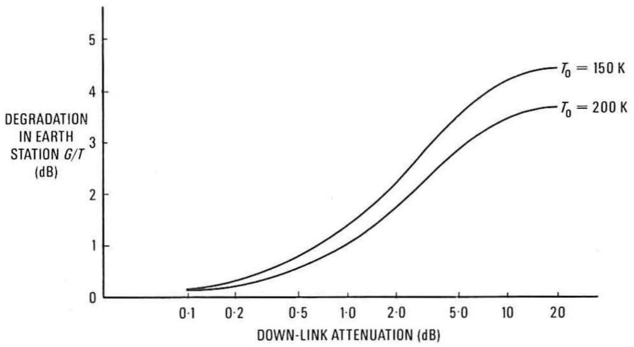



FIG. 3—Typical satellite TWT transfer characteristic




*Note*: Mean figures for Europe at 14 GHz

FIG. 4—Atmospheric attenuation against time percentage




$T_o$ : Total earth station receiver noise temperature in clear-sky conditions

FIG. 5—Degradation of earth station $G/T$ with increasing down-link attenuation

are usually those of worst months averaged over several years.

It is clear from equation (3) that $C/N_{tuo}$ is affected by up-link fading and $C/N_{tdo}$ by down-link fading. However, it should also be noted that up-link fades cause an increase in satellite input and, hence, output back-off, so that up-link fades also affect down-link $C/N_{tdo}$. The significance of this effect depends on whether the satellite TWT is being operated in the linear or non-linear region.

Where up-link fades are foreseen as a severe problem, they can be compensated for by providing for increasing earth station EIRP. This is termed *up-link power control.* One drawback, however, is that earth station transmitters have to be provided with an output capability higher than that required for most of the time. Another way of compensating fades, on both up- and down-links, is by earth station site diversity; that is, by providing an automatically-switched alternative earth station some ten or so kilometres away on the basis that very heavy rain affects only a small area at any one time. This, however, is a very expensive and complex option and is likely to be used only to compensate the very severe fading expected when use is made of the 20/30 GHz satellite bands.

Another effect of attenuation, as already mentioned, is that on the down-link, it causes a decrease in earth station $G/T$ because of an increase in sky noise. Fig. 5 indicates this relationship. Attenuation, therefore, has a double impact on $C/N_{tdo}$.

The other two important elements of equation (2) are $C/I_{uo}$ and $C/I_{do}$; that is, the elements that allow for interference. Interference to a satellite system arises from terrestrial radio-relay systems, from other satellite systems and internally from within the system. It is convenient to discuss each of these separately.

## Radio-Relay Interference

Much of the frequency spectrum allocated for the fixed satellite service by the World Administrative Radio Conference (WARC) is also allocated for use by terrestrial radio-relay systems. This applies in both the 6/4 GHz and 14/11 GHz bands. It means that agreement has to be reached on the operating parameters of each type of system so that excessive interference is not caused to the other, and the constraints that result from this agreement are the subject of CCIR Recommendations[10]. For satellite system design, two principal requirements result from these Recommendations. The first relates to total noise at the demodulator, where the designer is required to allow that 10% of total noise be caused by radio-relay interference. Most of this noise will originate on the down-link, and be caused by radio-relay transmissions being picked up by earth station antennas. The second requirement is a limit on the power flux density (that is, watts per square metre per 4 kHz) at

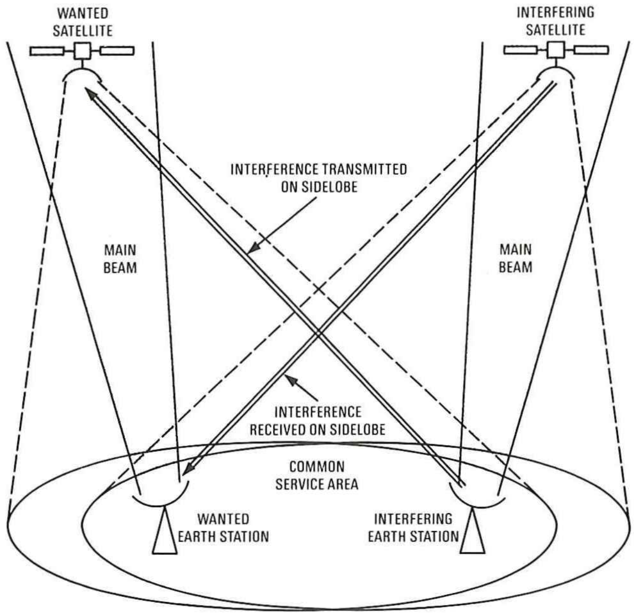



FIG. 6—Inter-system interference from earth station antenna side lobes

the earth's surface that can result from satellite transmissions. This results in a need for the application of energy dispersal signals at the transmitter to avoid excessive energy accumulation at certain spectral positions; it also limits to some extent the power that can be transmitted from the satellite.

These frequency sharing constraints will have to be considered when decisions are made on better performance requirements for satellites.

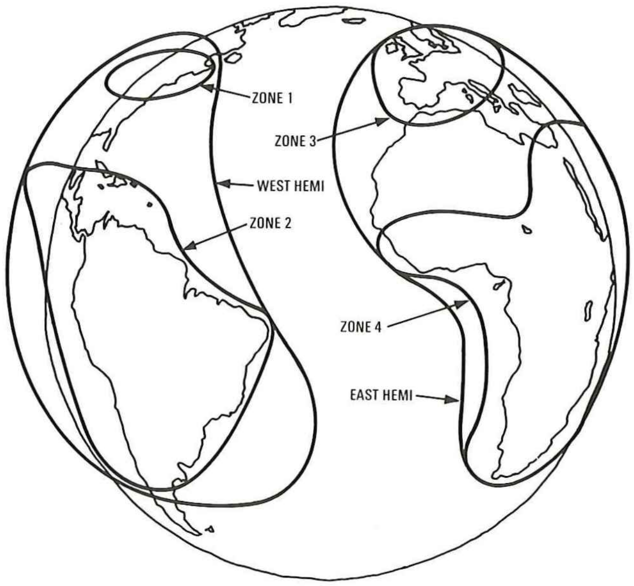
## Interference from Other Satellite Systems

Nearly all communication satellites are positioned in an earth synchronous (geosynchronous) orbit, and, because there is only one such orbit, there is an ever increasing need to locate satellites of different systems closer and closer together. The result is inter-system interference, resulting from energy from an adjacent system entering a wanted system by means of earth station antenna sidelobes, as shown in Fig. 6. This mechanism leads to the need for the lowest possible earth-station sidelobe levels (relative to main-beam gain) which, for up-links, tends to imply large antennas (that is, high gain) and low transmitter power. This unfortunately contradicts systems economics, which in general tend to favour small antennas. The alternative is to develop antennas with low sidelobe gains and, although this is now being achieved, it again has an effect on the economics. The level of inter-system interference that must be tolerated is the subject of another CCIR Recommendation[11]; the requirement in this case is that 15% or 20% (depending on frequency) of total noise at the demodulator input must be allowed for this mechanism. Because of the need to comply with limits of this sort, any new satellite system must be approved via a co-ordination procedure supervised by the International Frequency Registration Board of the ITU before it can be implemented.

## Internal System Interference

The two main mechanisms in this category are adjacent-channel interference (ACI) and co-channel interference (CCI).

ACI is the result of transmissions to one satellite transponder overlapping at band edge into the pass band of an adjacent (in frequency) transponder. It is caused by filter limitations coupled with the use of narrow guardbands




*Notes*: 1 The same frequencies are re-used in each of the six coverage areas
2 Zone beams use right-hand circular polarisation (on down-link)
3 Hemisheric beams use left-hand circular polarisation (on down-link)

FIG. 7—INTELSAT VI satellite beam coverage contours (6/4 GHz Atlantic Ocean Region)

between transponders, by transmission rates and modulation methods that require all the available transponder bandwidth, and by spectrum spreading caused by earth station/satellite TWT amplifiers operating near saturation. The result of reducing ACI by easing any of these contributory factors is a decrease in the capacity achieved from the total bandwidth available.

CCI results from further attempts to maximise capacity, in this case by frequency re-use. Frequencies are re-used in two ways: one by polarisation discrimination in which orthogonally polarised (linear or circular) transmissions at a common frequency are employed by different users; and the other by spatial discrimination (that is, the independent use of multiple satellite antenna beams each with limited coverage area). Fig. 7, which shows the beam coverage pattern for the INTELSAT VI satellite, illustrates use of both these techniques[12].

### TWT Effects

The other two elements of equation (2) are $C/N_{io}$ and $M$. Intermodulation results from the transmission of more than one carrier through the non-linear satellite TWT amplifier. As the operating point of the TWT is reduced from saturation, the transfer characteristic becomes more linear and intermodulation noise decreases. However, back-off also reduces satellite EIRP and so a balance has to be struck between reducing intermodulation noise and degrading down-link $C/N_0$. The result of intermodulation is that overall $C/N_0$ is reduced relative to the single-carrier case and, hence, the total multi-carrier capacity of a transponder is always less than single-carrier capacity. This is one of the main reasons for preferring TDMA to FDMA for high-capacity digital systems.

Another effect that results from TWT characteristics is a degradation of the transmitted signal. The normal modulation technique for satellite systems is PSK (2- or 4-phase) and the amplitude variations imparted to such transmissions by earth station transmit filtering, combined with the amplitude modulation/phase modulation conversion properties of the TWT, result in a closing of the eye pattern and a consequent degradation of demodulated BER. The increase in total $C/N_0$ necessary to compensate for this degradation is the margin $M$ in equation (2).

## CONCLUSION

Many diverse factors have to be taken into account in designing satellite systems to any particular performance requirements, and it is capacity and cost that have to bear the brunt of any significant improvement in performance.

There are, of course, many techniques available to the system designer which accommodate these factors and enable cost-effective systems to be designed. Some of these have already been mentioned in this article, including use of multiple spot beams, the choice between FDMA and TDMA, optimised satellite and earth-station parameters, some choice of modulation method (2- or 4-phase PSK) and use of forward error correction. Others that are likely to be available in the future include further, more spectrally-efficient modulation methods, lower-rate voice-encoding techniques, more attractive use of FDMA by employing satellite TWT linearisers and possibly on-board processing.

Unfortunately, lack of space does not permit a more detailed discussion of the application of these techniques. However, their existence, which results from the large amount of work constantly applied to optimising satellite systems, helps to ensure that the challenge of achieving high capacity at high quality and competitive prices will be met.

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### Biography

John Lewis studied at the University of Bradford from 1968 to 1972 and gained an honours degree in electrical engineering; he later obtained the Diploma of Imperial College in communications. After graduating, he joined British Telecom International, and has worked on many aspects of satellite system design. He is currently Head of the Digital Systems Group in the Satellite Systems Division. In 1979, he was appointed by CCIR to act as correspondent to CCITT SG XVIII with the responsibility of ensuring that due account was taken of satellite system requirements during development of ISDN error-performance objectives.

# Television Detector Vans

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UDC 621.397.62 : 621.396.677 : 629.114.7

*The Post Office has provided a service for the detection of unlicenced television receivers for more than 50 years. This article describes the vehicles and their detection equipment recently brought into service to replace those that have been operating since 1968.*

## INTRODUCTION

As part of its agency work in the collection of television licence fees for the Home Office, the Post Office provides a service to detect the use of television apparatus suspected of being unlicenced.

The detection service now spans over 50 years, being originally started in 1930 with a fleet of Morris 8 vans, which operated until 1939. The detection work was restarted in 1948 with smaller Morris vans, but these were replaced within five years by Hillman estate cars[1]. Ten years later, in 1962, a further change to Morris Oxford estate cars was made[2]. None of these vehicles was specifically designed for detection work, each one simply had the equipment installed in the rear and an aerial fixed to the roof.

In 1968, the Commer 2500 van was introduced into service as a purpose-built vehicle[3], having a raised roof containing the aerial control mechanism. It also enabled the detection officer to sit with adequate headroom at his equipment console situated behind the driver. The bodywork was lined against condensation, and hot air was directed at the windows to prevent misting. Even so, this was not sufficient to counteract the extremes of summer and winter weather that the staff had to endure for some 10 hours each working day.

These vans were successful and the 30 purchased were in service for many years, although diminishing in numbers recently, until they were finally replaced in October 1983 by the present fleet of 22 Leyland vehicles.

The detection vans are crewed by an enquiry officer, holding information on suspected premises, a driver, and an engineer to operate the detection equipment.

## DETECTION VEHICLE

A Leyland Sherpa van, with modified bodywork and additional battery power, is used. A raised roof and all-round body insulation with air conditioning provides a comfortable environment for the staff against the extremes of hot and cold weather. Windows cut into the raised roof section allow easy viewing when house markers are being set. A lockable cupboard unit is provided to secure both personal and official items.

Power for the detection equipment and air conditioning is supplied from an additional 12 V battery fitted in the engine compartment. A high-output alternator is provided to maintain the capacity of both batteries.

The detection unit is self-contained, with the bottom drawer containing its power units and the central drawer its control circuitry on printed-wiring boards. The remaining circuitry is housed in the desk-top units. (See Fig. 1.)

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FIG. 1—Interior of detection vehicle

## PRINCIPLE OF OPERATION

All modern television receivers use the superheterodyne system and have a local oscillator adjusted to 35 MHz above the required television signal frequency, and radiation from this unit could be detected from a distance exceeding 30 m. The range at which the signal can be detected has diminished considerably in the past 10 years because of the introduction of the solid-state tuner (which contains the oscillator), and this has had a noticeable effect on the performance of the old-type detector equipment. Compared with their predecessor, the new detectors have an increased sensitivity (some 10 times (20 dB) greater), and accurate detection to 50 m is possible.

Because of the screening possible with solid-state television tuners, the radiation patterns are very directional with extremely low levels of signal radiated in some directions. This makes the interpretation of the received information more difficult.

At the frequencies used (500–895 MHz), propagation of the radiated signal is affected by surrounding objects. In some cases, as a result of reflections, the received signal can be greatly reduced in level and appear to originate from an

FIG. 2—Detection vehicle showing aerial arrays

image located behind the reflecting surface. If the direct and reflected signals are received in the same phase, an enhanced signal is received. These variations make direction finding unreliable when a single aerial is used; hence, the need for two aerials.

Two connected and identical fixed aerials, adding in phase the received signals, produce a polar diagram that is the result of multiplying together the two expressions for their individual polar diagrams; this gives a multi-lobed pattern, which is dependent upon the spacing between the two aerials. Electronic adjustment maintains the lobe pattern for changes of frequencies.

The aerials used are broadband narrow-beam reflector type, each with four broadband elements and a panel reflector. Four of these aerials are mounted on the roof of the vehicle, two facing each side (see Fig. 2).

When the vehicle is driven past a signal source, the level of the received signal rises as each lobe points at the source. The distance that the vehicle travels between each maxima relates to the range of the source, and the relationship between the adjacent maxima can indicate the line at right-angles to the direction in which the vehicle is travelling along which the source lies (see Fig. 3).

For a reflected signal, the range indicates the distance between the vehicle and the image. This range is obviously going to be greater than that of the real source; so the signal

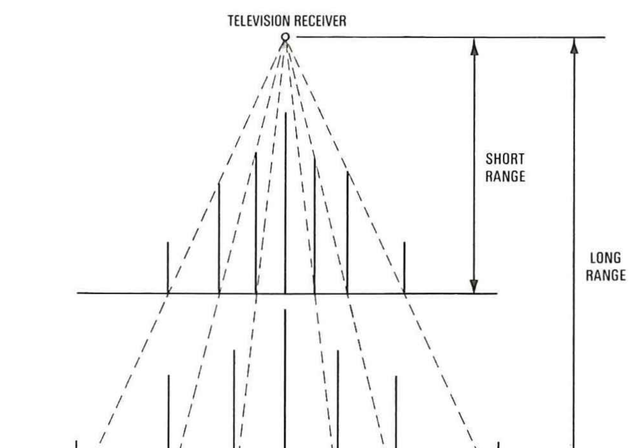


FIG. 3—Positions of maximum indications as aerial is moved past television receiver at long and short ranges

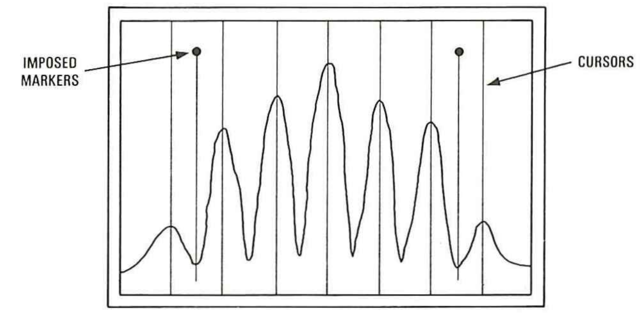


FIG. 4—CRT display of memorised amplitude against distance travelled

considered to be the shorter range is the real source, and constitutes the wanted information.

## PRESENTATION OF INFORMATION

A UHF radio receiver is used to locate the signal and display the incoming information. By sweeping its second local oscillator, a panoramic display of amplitude against frequency is displayed on a cathode-ray tube (CRT). To enable the required spacings between maxima to be measured, a solid-state random-access memory (RAM) is used. The RAM stores amplitude information, together with the output from a sensor that produces pulses as the vehicle moves.

After the signal source has been scanned, the RAM contains a display of signal amplitude against distance moved by the vehicle, in the form of vertical cursor lines. The result is a display of maxima and minima, which can be analysed by the operator to find the range and bearing of the source. (See Fig. 4)

The cursor is corrected for the change of frequency by means of a pre-programmed read-only memory (ROM). Two dots are added to the display by the operator as the vehicle passes the source, to indicate the boundaries of the premises thought to contain it.

## OVERALL FUNCTIONAL DESCRIPTION

A block diagram of the detection system is given in Fig. 5.

There are three main modes of operation of the detection equipment and these are described below.

### Panoramic Mode

The four aerials (two arrays) each have equal length feeders to the aerial switching unit. Six coaxial relays are used to select the aerials; these are controlled from the right-hand panel of the operating console (see Fig. 6) and enable a

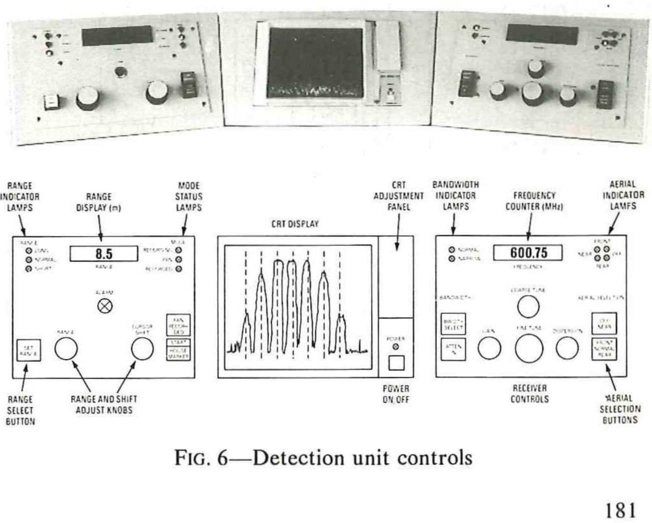


FIG. 6—Detection unit controls

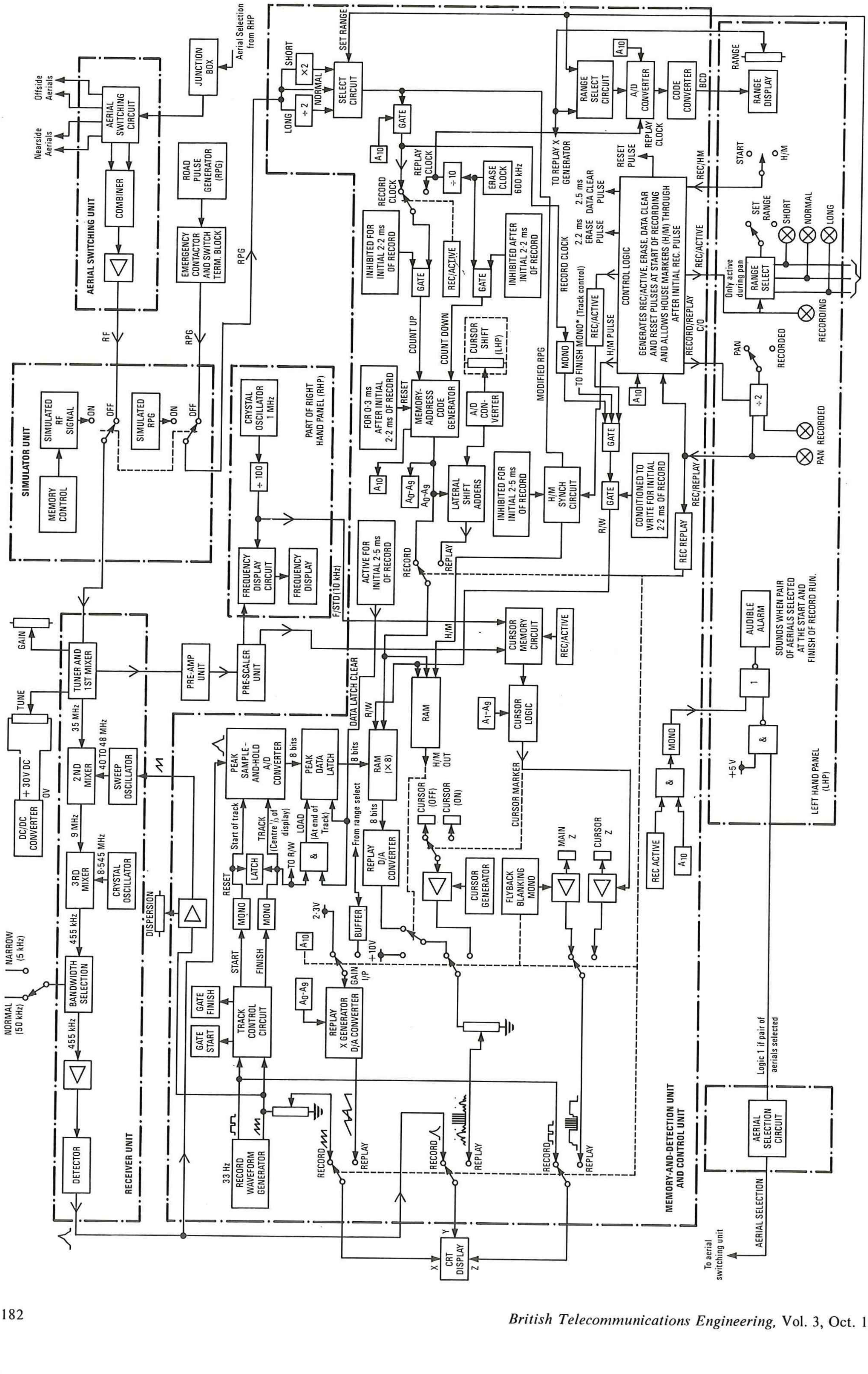


FIG. 5—Block diagram of overall detection system

choice to be made of either the front, rear or both aerials on one side of the vehicle. The output from the relays is either the signals from the two aerials, or the signal from one aerial and a 50 Ω termination. These signals are fed into a passive combining network; the result is passed to a pre-amplifier and then, via the OFF function of the simulator unit, to the receiver unit.

The input stage of the receiver is a television tuner, having a local oscillator output of 35 MHz. Tuning is provided and its power, together with that for the gain control, is obtained from the DC/DC converter board. The 35 MHz signal is amplified in the pre-amplifier, divided by 100 in the pre-scaler and fed to the offset frequency display, where the frequency of the received signal from the aerial is displayed in digital form. It is also fed to the cursor memory, to provide correct cursor markers for the frequency in use.

The second mixer (receive unit) is fed by the output from the tuner and a sweep oscillator; its voltage is obtained from the record waveform generator in the memory-and-detection unit; that is, the same source as the X-deflection for the CRT display in the PAN/RECORDING mode. This control voltage, at minimum dispersion, causes the oscillator to sweep from 40 to 48 MHz. The second mixer is tuned to 9 MHz, so that the display centre shows the signal at the indicated frequency on the digital frequency meter. Altering the DISPERSION control reduces the control voltage amplitude and thereby reduces the range of frequency sweep of the oscillator, and in turn that of the display.

The third mixer converts the 9 MHz signal to 455 kHz before it is conditioned via a selector circuit to either a 5 kHz or 50 kHz bandwidth. The signal then passes via a pre-set amplifier and detector to provide, at a maximum of 9·1 V:

(*a*) the Y-deflection signal for the CRT display in the PAN/RECORDING mode, and
(*b*) the input to the peak sample-and-hold circuit in the memory-and-detection unit.

The record waveform generator provides:

(*a*) a sawtooth waveform for the dispersion control amplifier (modified for receiver sweep control),
(*b*) a sawtooth waveform for the track control circuit,
(*c*) a sawtooth waveform, for the X-dispersion signal to the CRT in PAN/RECORDING mode,
(*d*) blanking pulse output for the Z-deflection signal to the CRT in PAN/RECORDING mode, to eliminate display during flyback, and
(*e*) blanking pulse input to the track control circuit.

Therefore, in this PANORAMIC mode, the CRT displays signal amplitude against frequency, the range being dependent on the DISPERSION control setting. The digital frequency meter indicates the frequency at the centre of the CRT display (altered by the TUNE controls). The amplitude of signals on the display is altered by the GAIN control.

The ATTENUATOR switch inserts a 40 dB attenuator into the radio-frequency (RF) signal path within the aerial switching unit, between the combiner and the amplifier.

The RANGE control can be set to SHORT, NORMAL or LONG, depending on the length of road run to be used during recording; this depends on the distance of the television receiver from the vehicle. This setting also modifies the limits as indicated on the digital range display (in REPLAY mode only).

## Recording Mode

### *Initiation (First 2·5 ms)*

When a television oscillator signal has been located, and a pair of aerials selected, the vehicle is positioned for maximum response, and the GAIN control is adjusted until limiting of the signal is just apparent on the CRT. The range of the television receiver from the vehicle is then assessed, and the set range selected (set range is active only in the PANORAMIC mode). The ranges are:

(*a*) SHORT—2 to 11·5 m,
(*b*) NORMAL—2·5 to 22·5 m, and
(*c*) LONG—3·5 to 45·5 m.

At the start of a road run, the START/HOUSE MARKER button is pressed to initiate the RECORDING mode (*rec/active* signal generated).

For the first 2·2 ms, an *erase* pulse generated in the control logic ensures that any earlier information is erased.

(*a*) The record clock, derived from the road pulse generator (RPG) is prevented from activating the count-up input of the memory-address code generator.
(*b*) The erase clock, generating in excess of 1024 pulses (full memory) in this erase time,is connected to the count-down input of the memory-address code generator.
(*c*) The peak data latch in the memory-and-detection unit is cleared of data and the load function gated.
(*d*) The read/write (R/W) input to the 9 RAMs constituting the data memory is held in the WRITE condition.

As soon as address $A_{10}$ is reached by the memory-address code generator, an *alarm out* signal is generated from the control unit. If a pair of aerials has been selected, the audible alarm sounds, to signify that the run is satisfactory. If only one aerial has been selected, no alarm sounds. For 0·3 ms following ERASE, RESET mode is generated from the 2·5 ms *data clear* pulse in the control logic to reset the memory-address code generator, while the peak data latch is still prevented from loading.

From the start of recording, the *rec/active* signal illuminates the RECORDING light-emitting diode (LED) indicator.

### *Recording*

The RPG consists of a proximity detector mounted on the vehicle differential gearbox casing, and detects passing teeth on a disc mounted on the propellor shaft to provide a signal relative to vehicle movement. These signals are fed, via the OFF function of the simulator unit, to the range-selector circuit where the pulses are conditioned. This conditioning depends on the range selected, and ensures that, irrespective of distance travelled, the 1024 bits required to fill the memory produce a full detection pattern of received lobes. This is achieved as follows:

(*a*) LONG selected—pulses from RPG divided by 2.
(*b*) NORMAL selected—pulses fed through the circuit.
(*c*) SHORT selected—pulses from RPG multiplied by 2.

When 1024 pulses have been received and the memory is full, an $A_{10}$ lockout circuit blocks the RPG pulses to ensure that the first part of the memory is not overwritten and the data lost. The pulses are then fed, via the initial 2·2 ms erase inhibit gate, to the memory-address code generator. These pulses represent the record clock.

The output from the address code generator is fed to the 9 RAM data memory. Only outputs $A_0$–$A_9$ are used; $A_{10}$ is used as a timing control. The RPG pulses also generate the write signals to the data memory, which consists of eight 1024 × 1 bit arrays, for peak data, and an extra array for the house marker.

The record waveform generator feeds output ramp and blanking pulse signals to the track control unit, which modifies them to provide the inputs to the peak sample-and-hold plus peak data latch circuits:

(*a*) *Reset* pulse Present at the start of track, that is, at the start of the centre third of the ramp, and, therefore, the display sweep.
(*b*) *Track* pulse Present for the duration of the centre third of the display sweep.

(*c*) *Load data* pulse  Present at the end of track; that is, at the end of the centre third of the display sweep.

The analogue signal from the receiver unit, fed to the Y-input of the display, is also fed to a peak detecting sample-and-hold circuit in the memory-and-detection unit, which, owing to the *reset* and *track* pulses, operates on the middle third of the sweep, holding the maximum analogue value in this area and converting it to an 8 bit digital word. At the end of each track, the 8 bit word is loaded into the peak data latch. This peak reading is written into memory when the track is complete and an RPG pulse has been received. At the start of the next track, the peak detector is reset to accept the next centre third of sweep peak. Fig. 7 shows the relative timing of these pulses.

Depression of the START/HOUSE MARKER button indicates house boundaries during the record run, These pulses, being asynchronous, do not necessarily occur at an exact store address, and are held in a house-marker synchronising circuit, until a road pulse is generated. The output signal, inhibited during ERASE mode, is passed to the house-marker memory.

The cursor memory circuit is pre-programmed with information about the position of the cursor lines. Two input signals are required: a standard frequency signal, derived from the crystal oscillator, and the tuning signal derived from the pre-scalar unit. This circuit is active only whilst a *rec/active* pulse is present.

During the recording period, the display indicates amplitude of signal related to frequency and, as the maxima and minima of received signal levels are passed, the displayed signal rises and falls.

Recording is ended and REPLAY mode is entered either

(*a*) automatically when 1024 record clock pulses have been received and address $A_{10}$ is active from the memory-address code generator, or

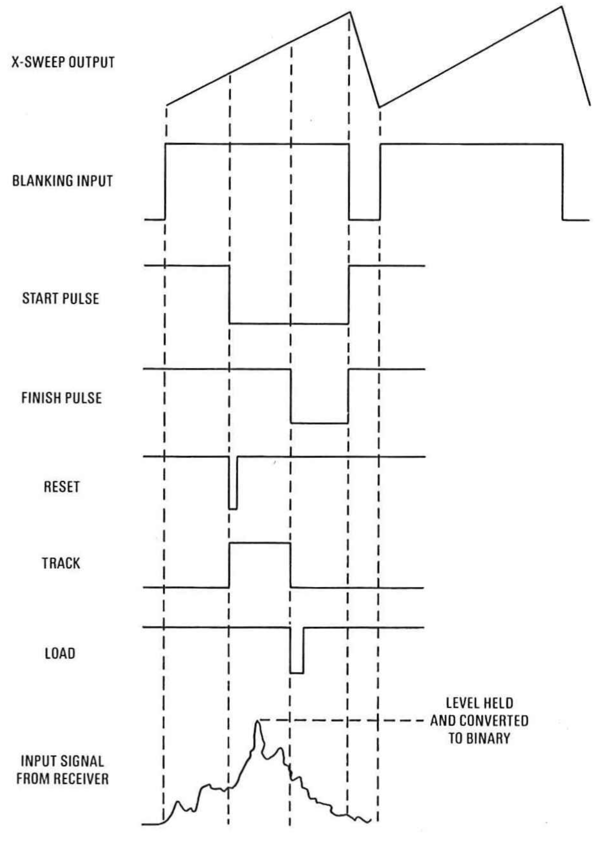


FIG. 7—X-waveform generator, track control and peak detector circuit waveforms

(*b*) by the operator depressing the PAN/RECORDED button, if a full run is not completed.

In the case of (*a*) the address $A_{10}$ signal:

(*a*) inhibits record clock pulses to the address code generator,

(*b*) generates a *record/replay* change-over signal, and

(*c*) provides an *alarm out* signal, which will sound the audible alarm, to signify the completion of a good run.

The *record/replay* change-over signal causes the divide-by-2 circuit to change from PAN to RECORDED state and the LED to light. The *rec/replay* signal, by changing to REPLAY, provides the following functions:

(*a*) change-over of the X, Y and Z-inputs of the CRT to the replay lines,

(*b*) introduction into the address lines of the memories active lateral shift adders for the cursor shift control,

(*c*) removal of the *rec/active* signal in the control logic causing:

(*i*) the memory-address code generator to be fed from the replay clock (the erase clock divided by 2),

(*ii*) the read/write circuit to provide a read-only signal to the memories,

(*iii*) latching of the cursor memory circuit, and

(*iv*) the RECORDING LED to be extinguished.

The CRT now displays the output of the memory circuits; that is, signal amplitude relative to distance travelled by the vehicle.

When a recording is terminated by the PAN/RECORDED switch being depressed, the memory is not full and the address $A_{10}$ signal is not present. The alarm does not sound, and the *record/replay* change-over signal is not generated, although depressing the switch carries out this function on the divide-by-2 circuit and changes the output to a *replay* signal.

## Replay Mode

The replay clock cycles the memory at approximately 67 Hz and the following information is required on the CRT during alternate sweeps of the time base.

(*a*) Information generated by the cycled address; that is, the amplitude display of the stored peak data, plus house markers, together with a means of aligning this pattern with the cursor.

(*b*) Sweeps, to display cursor lines, derived from the cursor memory circuit. The outer three lines on each side are adjustable to enable ranging of the amplitude pattern.

These two displays appear superimposed on the CRT display.

The CRT X-deflection signal is obtained from the replay X-generator, via a 2-pole switch. The range control provides a voltage, via a buffer amplifier, to one input of the switch; the other input is a reference voltage. The address code generator signal $A_{10}$ operates the switch, causing it to change-over every time this generator completes a cycle. The reference voltage is fed to a multiplying digital-to-analogue (D/A) converter. The cycled address, $A_0$ to $A_9$ is also fed to the D/A converter. The effect of the two voltages to the converter is to produce alternate X-deflection ramps, one of constant shape, the other a variable slope dependent on the setting of the RANGE control. The recorded analogue peak signal information is supplied to the CRT only during the time of the constant slope ramp (see Fig. 8).

The CRT Y-deflection signal, during the constant-ramp X-signal, is obtained from the 9 RAMs forming the data memory. At each address with a house marker stored, the Y-input is taken from the main memory output and presented with a +10 V offset signal, providing a dot at the top of the display above the maximum level of the amplitude

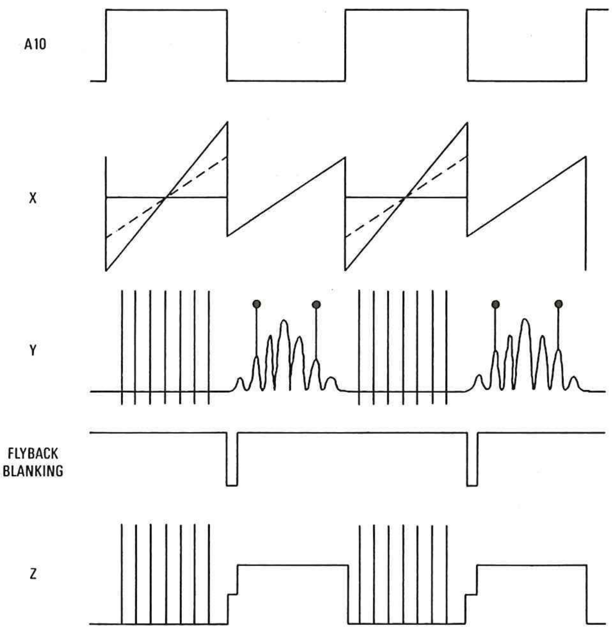


FIG. 8—X, Y, and Z replay waveforms

pattern. The cursor control provides a voltage to the memory-and-detection unit, where it is converted to a digital word and added to the address word. As this may be larger or smaller than the current address word, it therefore changes the address up or down, without its continuously cycling nature being affected. The effect on the displayed data information is to move it left or right to align the pattern with the cursor (see Fig. 8).

On the variable ramp sweeps of the replay X-generator, cursor signals are fed to the Y-deflection input. These marker pulses are generated in the cursor logic circuitry from the address code and the cursory memory codes, $D_0$–$D_7$, to operate a switch, which alters the output level of an amplifier fed with the triangular signal from the cursor line generator. The amplifier output provides a visible display for cursor line positions and deflects the remainder of the output of the cursor line generator off the screen of the display. Therefore, the RANGE control can position the vertical cursor lines to fit the analogue pattern produced during the period of the constant slope ramp. (See Fig. 8).

Two Z amplifiers, one for each alternate sweep of the X-deflection signal, provide the brightness modulation on the Z-deflection on the CRT. The Z-modulation signal over the constant slope ramp is a blanking pulse at the transition of $A_{10}$, plus a normal brightness control for the display of the recorded data.

Cursor modulation signals are produced from the cursor marker information, and provide additional brightness for the otherwise feint cursor lines (see Fig. 8).

Depending on the range control selected, output voltages are modified in a selector circuit, converted to binary and used to derive the inputs of a digital range display. Once the operator has completed the match between pattern and cursors, this display indicates the range of the television receiver from the position at the centre of the vehicle recording run.

## USE OF THE SIMULATOR UNIT

A simulator unit within the centre drawer of the detection console enables the equipment to be operated while the vehicle is stationary and without an external RF signal source. A pulse generator simulates road pulses and a UHF oscillator, by means of a voltage-controlled attenuator and a pre-programmed memory, provides a varying amplitude of signal as an idealised version of that received from a real source when the vehicle is moving. The simulator, set to ON, provides a static demonstration and a check on the correct functioning of the equipment, with the exception of

(*a*) the roof aerials,
(*b*) the aerial switching unit, and
(*c*) the road pulse generator.

The output signal from the simulator is at approximately 600 MHz, and the simulated range in REPLAY on all ranges is between 9 to 9·5 m. This signal is not intended to assess system sensitivity. The fixed RPG rate and conditioning input circuits in the control unit are controlled by the selected range; the recorded patterns appear as in Fig. 9.

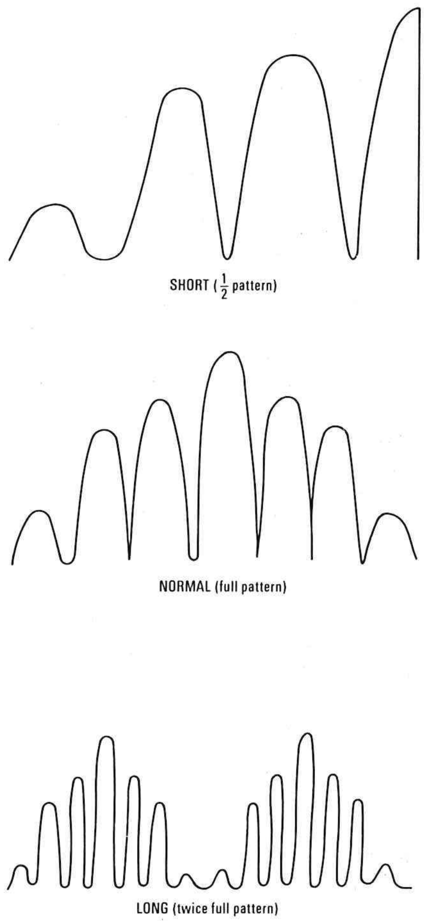


FIG. 9—Recorded patterns from the simulator unit

## OPERATIONAL CONTROL

A Post Office Headquarters Department known as *National Television Licence Records Office* (NTVLRO) located in Bristol is the centre for the planning and operational control of this agency service. With the records based on postcoded addresses, it is relatively simple for the NTVLRO to check through computer records and ascertain the type of licence, if any, credited to a particular address

Most of the NTVLRO's work is recording licences purchased over the counter at a Post Office or by application through the post, but it is the work of one section to check for possible licence evasion. When an area of the country shows a tendency towards higher than average evasion figures, appropriate measures are initiated.

The detection vehicles are headquartered regionally and an annual programme prepared for each van defines the areas of the region where it will operate each week. The local area enquiry officer, who has information on where

unlicenced sets are suspected of being operated, arranges for detection runs when the sets are most likely to be switched on. The vans also comb an area by travelling along roads where unlicenced premises exist. During these combs the engineer looks for signals on the equipment from all nearby buildings. If an unlicenced property appears to be generating a signal, then a complete check is made.

The Post Office, through the NTVLRO, also controls any subsequent action, including prosecution, that may be initiated when unlicenced apparatus is detected. To assist when giving evidence in such cases, it is customary for the engineer and the enquiry officer to accompany each other when dealing with the public so that each person can confirm the words and actions of his colleague. The driver is instructed to remain with the van to deter vandalism.

Periodically, the normal weekly work is broken by a national campaign, when the detector vans from various regions are grouped into a fleet comprising eight or more vans. This fleet then works in an area for possibly three weeks, moving on to other areas for similar periods until a region has been thoroughly checked.

There is no secret about this anti-evasion work, the whole operation is given wide local publicity so that persons have an opportunity to purchase a licence before detection checks are made. The vans, with their prominent aerials, are clearly marked *Television Detector*. Prior to visiting an area, the local newspaper is asked to print a photograph of the vans together with a description of its capabilities and the date of the proposed visit. During the visit the van is driven to busy places such as markets, schools and factories to give a maximum display of the vehicle, and local television channels occasionally mention the whereabouts of the vans.

The engineers are recruited as volunteers for a two year tour of duty, but many request to stay on for much longer periods, finding the work interesting and their contact with the public and the authorities very stimulating. Reaction from people challenged about possession of a licence can vary in young and old, men or women, from tears at the thought of the publicity and prosecution, to extreme aggressiveness at being disturbed just to be asked such a question.

## CONCLUSION

The present system of detection represents an improvement on the previous equipment by using up-to-date technology to give a greater sensitivity and effectiveness over a larger distance from the van. Eight vans have been in use since October 1983 and 20 vans since January 1984. High reliability is expected from the equipment with very little time lost due to maintenance. Apart from some early minor faults all the vehicles are functioning as expected. Of the 22 vans in the fleet, two vans are used on secondary programmes which enable them to be used as replacements in the event of accidents or breakdowns.

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## Biographies

K. Martin served in the Fleet Air Arm from 1956–1970, becoming Chief REA in charge of the radar and radio workshop. He joined Kenure Developments Ltd. in 1970 as Technical Manager associated with the design and development of overall test facilities for guided weapons. He has also been involved with test-set designs covering microprocessors, uncommitted logic arrays, CMOS, TTL and various pneumatic and hydraulic interface controls for the Ministry of Defence, British Aerospace and other manufacturers of defence systems. He has recently been involved with monitoring and metering systems for British Telecom. He became Technical Director of Kenure Developments in 1978.

S. A. L. Ward joined the Post Office in 1941 and worked on 2000-type director exchanges. After four years as an instructor on telegraph machines and line transmission equipment at the Technical Training College, he was promoted in 1956 to the Engineer-in-Chief's office to work on the development of customer apparatus. He moved to the Postal Engineering Branch in 1969 on project work concerned with building mechanised parcels offices. In 1975, he became a Head of Group employed on overseas consultancy work before moving to the Post Office security and special projects group.

# Design and Performance of Digital Transmission Systems Operating on Metallic Pairs

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UDC 621.372.2 : 621.374 : 621.395.49

*Consideration is given to the design and performance of digital transmission systems operating over metallic cables for use in different parts of the telecommunication network. Particular attention is given to thermal noise, crosstalk, and impulsive noise limited systems in the context of achieving acceptable error performance. A methodology of performance specification is discussed and the importance of providing immunity against so-called network effects is emphasised. The status of international studies with respect to the performance specification of digital line systems is reviewed. The actual error performance of some operational and field-trial systems is reported, and error clustering is found to be significant on some systems.*

*This article is a revised version of a paper which first appeared in The Radio and Electronic Engineer and is reproduced here by permission of the Institution of Electronic and Radio Engineers**. This updated version takes due account of the final CCITT SG XVIII meeting during the current study period (1980–1984).*

## INTRODUCTION

Metallic pair cables have been used since the advent of telecommunication networks to carry the majority of traffic at all levels of these networks. Because of the quantities of cable that have been installed, this will remain for many years to come. Most modern systems are characterised by a large number of dependent repeaters, spaced about 2 km apart, so that the signal does not suffer undue attenuation and become vulnerable to noise.

For digital systems, the performance is determined by regenerative repeaters that amplify, retime and regenerate the digit stream. Pulse distortion caused by the metallic pair is equalised so that the pulse spectrum at the regenerator 'eye' meets Nyquist's first criterion and thus avoids intersymbol interference. Fig. 1 illustrates that the pulse spectrum shaping (roll-off) can be chosen from many possibilities with differing noise immunities that can be determined by computer calculation or simulation. The best roll-off for one type of noise is not necessarily the most suitable for another, as their spectra may differ, and there may be a need to compromise. Although a steep spectrum roll-off tends to improve performance for all sources that are dominated by high frequencies in the equalised spectrum, jitter in the extracted clock tends to be worsened by steep roll-offs. General improvements are obtainable by using a line code with minimised baud rate; for example, British Telecom (BT) is adopting MS43 for 2 Mbit/s systems on balanced symmetric pair cable and 6B-4T for 140 Mbit/s systems on coaxial cable.

Routes between nearby local exchanges are generally served by cables of quads laid up in layers, or pairs laid up in units. On these routes, a pulse-code modulation (PCM) carrier is now often used with 24, 30 or 48 one-way channels on each pair. The regenerators tend to be sited at points

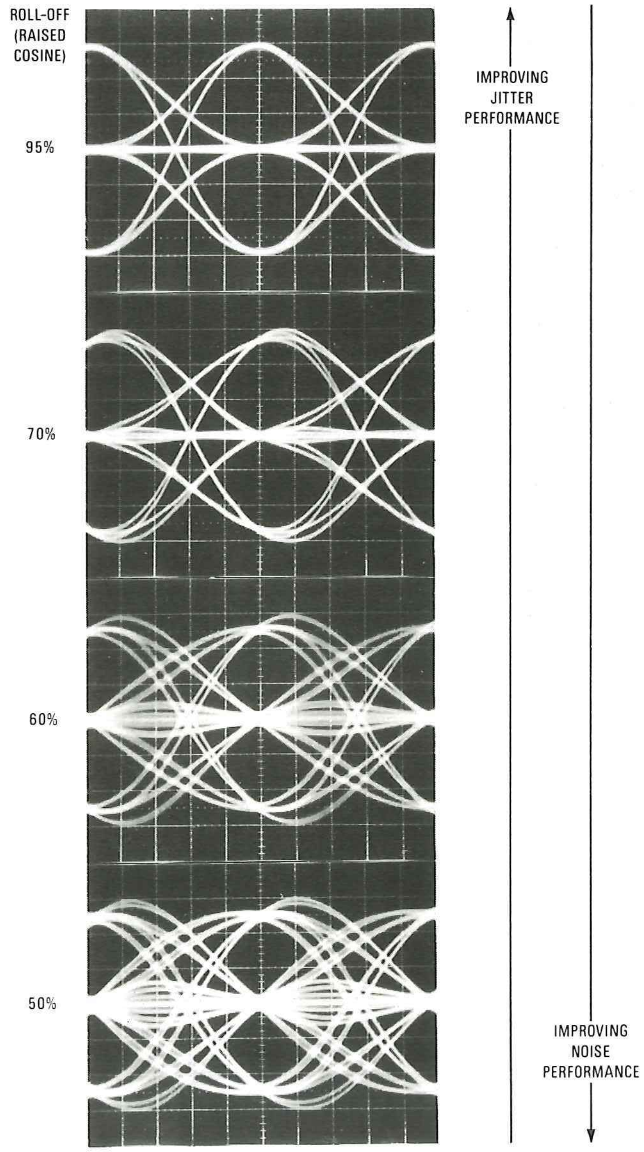


FIG. 1—Repeater eye patterns and performance trade-offs

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** McELVANNEY, S. G., NORMAN, P., and CATCHPOLE, R. J. Design and performance of digital transmission systems operating on metallic pairs. *The Radio and Electronic Eng.*, Apr. 1984, **54**(4), pp. 199–209.

where loading coils were previously housed; that is, within about 0·9 km of each end and with about 1·8 km spacing in between. The number of channels that can be carried is limited mainly by crosstalk from other systems sharing the same cable and, in particular, by near-end crosstalk when opposing directions of transmission are used within the same cable[1]. Far-end crosstalk is predominant when separate cables are used for the two directions of transmission, and more complex mechanisms, such as third-circuit crosstalk, are limiting in some situations. The crosstalk produced is dependent on the pattern content of the transmitted signals, and there is now interest in using scramblers to avoid particular patterns that result in high crosstalk, such as the all-ones *alarm-indication* signal, and to decrease the time variability of crosstalk. As digital switching is implemented, systems operating in the same cable will work from a common clock in a synchronous manner; such operation tends to increase the effects of crosstalk owing to amplitude rather than power addition of the interferers, and further study and field measurements are required to quantify its influence.

For high traffic densities on the trunk routes, coaxial pairs are used with inner/outer diameters of 0·7/2·9 mm, 1·2/4·4 mm or 2·6/9·5 mm. The larger the diameter, the lower the attenuation of any given signal per unit length, so the large diameters are generally used for the high bit rate system offering large channel capacities. The repeater spacings are usually made compatible with those of the existing analogue systems, so that repeater locations and underground housings can be co-sited. For example, the repeater spacings for the 140 Mbit/s system using 1·2/4·4 mm diameter coaxial cable are nominally 2 km, the same as for 12 MHz analogue systems. On coaxial cable systems, repeater spacings are determined by thermal noise. The crosstalk performance of these cables is adequate, but, with bit rates above 140 Mbit/s, reflections due to impedance discontinuities can cause significant degradations in system performance, and re-work on joints and terminations of existing installed cable may be necessary to ensure satisfactory performance[2].

The number of channels carried, and hence the bit rate, affect the circuit technology, repeater spacing and cost. In general, the cost per channel per kilometre reduces as the bit rate increases. However, at some point, the cost of the technology required to engineer the system causes the cost per channel per kilometre to start increasing.

At present, twisted symmetric pairs of copper or aluminium wire are generally used for access by each customer to his local exchange. In order to give a customer digital access to the integrated digital network (IDN) or integrated services digital network (ISDN), systems for transmission on one or two pairs, at rates of 80 kbit/s or higher, are beginning to be introduced in this part of the network. Furthermore, pair saving is being achieved to cope with the increase in the number of customers by the introduction of digital multiplex of 10, 24 or 30 channels.

## PERFORMANCE REQUIREMENTS

Design procedures for transmission systems are based, on the one hand, on international recommendations for overall system performance and, on the other hand, on the requirements of the local administration on size, power consumption, environment and reliability. In metallic-pair systems, there has to be many dependent regenerators at relatively inaccessible sites, for which the performance and local administration requirements are particularly severe. A brief review of the key factors is given here.

It is vital to ensure adequate operating margins in the repeater design to allow for imperfections under extreme conditions. A summary of key impairments is given in Table 1. To determine how large this margin must be, the shape of the eye pattern and noise-power degradations for each imperfection must be estimated[3].

### Noise Assumed for Design

Errors can be caused by inter-system crosstalk, thermal noise, external interference or faults, and the design must ensure that their magnitude does not exceed a specified value. Much work is being undertaken by the CEPT† and the CCITT* to derive satisfactory performance objectives, and these are reported later in this article. See also Reference 4.

Although the detailed specification of repeater performance is not expected to be a subject for international agreement, the method of measurement may be. It is essential to measure the operating noise margin of repeaters during manufacture to monitor their quality.

The noise margin in a symmetric-pair repeater may be expressed, for design purposes, in terms of a *repeater crosstalk noise figure*. There are separate expressions for near-end crosstalk and far-end crosstalk; and it has been proposed that these form Annex A to CCITT Recommendation G917[5]. Care has to be taken with this method when comparing alternative designs that carry the same traffic bit rates at differing line rates; the reference rate should be taken as that for the system interface. Nevertheless, this is a breakthrough in the achievement of a common basis for specification. In the application of a repeater crosstalk noise figure to the achievement of design objectives, the importance of repeater spacing, cable characteristics, effects of synchronism, jitter and installation planning factors, must

† CEPT—European Conference of Posts and Telecommunication Administrations

* CCITT—International Telegraph and Telephone Consultative Committee

**TABLE 1**
**Sources of Repeater Margin Impairments and Test Equipment Inaccuracies**

| Type of Impairment | Repeater Imperfections | Test Equipment Inaccuracies | Field Degradations |
|---|---|---|---|
| Eye closure | Equaliser/ALBO response<br>Equaliser/ALBO threshold<br>Amplifier gain and shaping | Cable simulator | Cable types<br>Structural echo<br>Temperature<br>Output of previous repeater<br>Ageing<br>Incipient faults |
| Noise enhancements | Data thresholds<br>Sample pulse width<br>Sample pulse timing | Any noise shaping filter<br>Noise Gaussianity<br>Noise meter bandwidth<br>Calibration | External interference surges |
| Jitter | Clock thresholds<br>Clock timing | Pattern dependency | Temperature<br>Ageing |

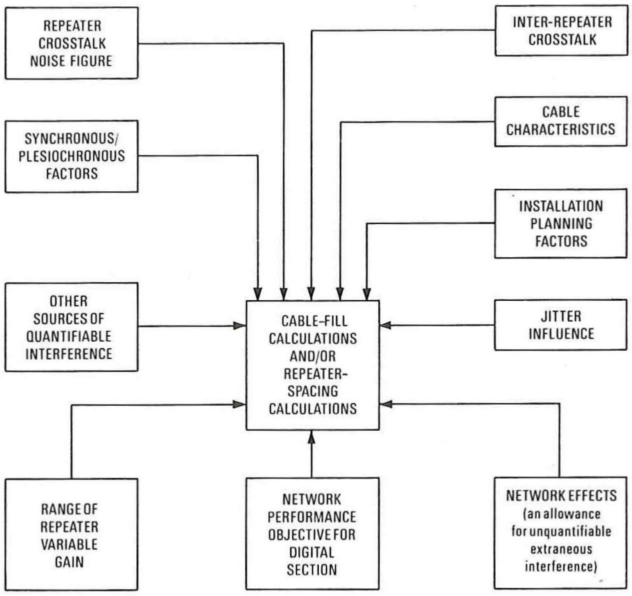



FIG. 2—Factors impacting on the error performance of a digital line system (crosstalk-limited systems only)

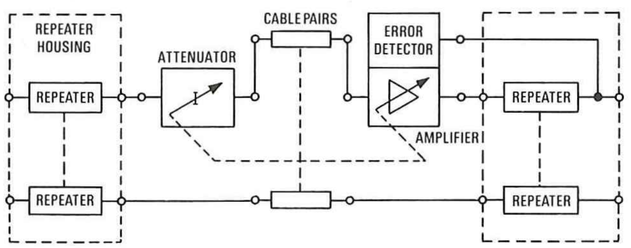



FIG. 3—Repeater field margin measurement

be considered as shown in Fig. 2.

It is important to have a capability to measure system noise margins. A simple, but effective, technique for use with symmetric-pair systems is illustrated in Fig. 3. The source signal is attenuated before being passed to line and, after traversing the line, is amplified by an equal amount before going to the regenerator under test. In this process, all external noise is amplified, including any crosstalk from other working systems in the cable. The degree of attenuation and amplification to produce the target error performance is a direct measure of the noise margin for that repeater section.

Similarly, the noise margin in a coaxial-pair repeater can follow the internationally agreed definition and measurement methods for *repeater noise margin* that have been proposed to form Annex A to CCITT Recommendation G918. In the recommended 'Method A', the amount that the noise level at the repeater decision point has to be increased to give a specified error ratio is measured. The repeater is operated over a section of given attenuation, and the additional noise is injected at the repeater input. In the alternative, 'Method B', the applied noise at the input that causes the given error ratio is measured directly. Method A relates more directly to the definition of noise margin, but access to the decision point inside the repeater is necessary and the automatic gain control must be manually held at the value for normal operation so that the gain does not change when the signal is removed during the noise measurements. Method B has the disadvantage that the noise levels at the decision point must be evaluated by means of the transfer function and noise figure of the pre-amplifier/equaliser, which may vary between repeater samples.

## External Interference

External interference due to natural and man-made sources must be considered, as it can cause errors to be introduced. Error-performance objectives for installed systems need a substantial allowance for such unquantifiable extraneous interference over and above the noise assumed in the design. This aspect is further discussed later in this article.

Extreme levels of external interference, due to lightning or faults in electrical power distribution systems, can even cause damage to the cable and repeaters. Obviously, protection against direct lightning strikes is impracticable, but a considerable degree of surge protection can be provided by using gas-tube spark gaps at the repeater input and output ports, backed up by semiconductor diodes provided within the repeater.

Certain tests to check the effectiveness of the surge-protection arrangements at repeaters are recommended by the CCITT. These involve the application of pulses of defined shape and amplitude (up to 5 kV) to the input and output ports[6]. This is a great help to the equipment designer since it defines the degree of protection that has proved to be adequate in practice. Isolation of the coaxial-cable outer conductor from ground reduces the magnitude of any induced surges, but is less safe for maintenance personnel.

Interference at levels that can cause errors can also arise from electrical machinery, switching, signals on pairs or quads in the same composite cable, and other analogue or digital repeaters installed in the same buried housings. This interference is reduced by the screening of the cable pairs and repeaters. The screening provided on the types of coaxial pair recommended by the CCITT is very good at high frequency (for example, crosstalk between pairs is greater than 140 dB from 1 to 1000 MHz), but low frequencies can penetrate and sometimes cause errors. It is, therefore, an advantage to make the low-frequency cut-off point of the repeater input circuits as high as possible consistent with satisfactory operation. The pulse droop resulting from this can be compensated by using the technique of quantised feedback.

The repeater can, of course, be screened as effectively as desired by suitable design of its case, although this may be expensive. It would be very useful for the designer if recommended tests for external interference immunity could be devised, analogous to those for surge protection and based on a consensus of the experience of operating companies and administrations. Published data on the error performance of installed digital transmission systems, and the identification of the sources of interference causing errors, are very scarce at present. System designers are awaiting the quantification of these effects.

## Jitter

The magnitude of any jitter present must be limited to enable any system to interconnect with other digital equipment, such as multiplexers, without producing errors, and to avoid the degradation of certain types of digital signal, such as television. However, the overall jitter performance of systems can be controlled by the use of jitter reducers at the receive terminals; in practical designs, these devices are often provided.

Also important is the control of jitter within a repeater (alignment jitter), so as to avoid a significant reduction of margin that may be caused by sampling the incoming pulses at the incorrect instants. Also, high-frequency jitter generation in the repeater after the decision point, as a result of internal crosstalk etc., can cause margin reduction in the next repeater along the line, because of the inability of its extracted clock to track this jitter.

Jitter is discussed further in an article in an earlier issue of this *Journal*[7].

## Transmission Delay

Each regenerator introduces a delay of less than one bit to the signal, which propagates over the metallic pair at a speed close to that of light. Therefore, the total delay of a system, although proportional to its length, is not usually significant.

## Power Feeding

Power feeding of intermediate repeaters along the cable conductors is necessary as reliable supplies are not otherwise available. Thus, to obtain a large distance between the power-feeding stations, the circuit must be designed to use the minimum of power. For the safety of installation and maintenance personnel, voltage and current amplitudes are usually limited. For example, in the UK, BT specifies a maximum of 250 V DC and 50 mA constant current on the inner conductors of coaxial cable systems, and a maximum of 75 V DC on symmetric-pair systems, unless rather special complex safety measures are made.

## Supervision

As coaxial systems use many unattended repeaters, it is necessary to monitor their performance so that any degradations or faults can be located rapidly and rectified.

A most useful characteristic to monitor is the in-service error performance; by using a separate telemetry channel (for example, on the coaxial pair at frequencies below the traffic signal) information can be conveyed to attended terminals. The arrangements should be able to locate intermittent faults; consequently, if the repeaters are interrogated sequentially, an error store is needed at the repeater site as the fault may occur between interrogations.

Cable faults can interrupt the power-feeding current and disable the telemetry paths. In this case, a DC method of fault location can be used. At each repeater, a high-value resistor in series with a diode is connected from the power supply to the coaxial outer conductor so that the diode is normally non-conducting. To locate a break in the cable, the polarity of the power feeding is reversed and the value of the current at the terminal station indicates the location of the fault.

## Reliability and Lifetime

The large number of regenerators in a typical metallic-pair system, and the stringent network-availability requirements for the system, result in a tough reliability requirement on each repeater. Mean-time-between-failure (MTBF) requirements of several hundred years are commonplace to ensure that repeater replacements are not required by an administration on a daily basis! Modern circuit technology allows the achievement of such figures, even under the wide range of environments encountered in a network. The correct choice can be different for different systems; for example, discrete components on printed circuits may be satisfactory for 140 Mbit/s systems, whereas hybrid techniques may be necessary at 565 Mbit/s in order to achieve the required circuit speed. A remarkable feature of telecommunication equipment is the common requirement for an assured lifetime of 25 years, although this may lessen with the quickening pace of technological obsolescence.

# EQUIPMENT DESIGN OBJECTIVES IN TERMS OF ERRORS FOR DIGITAL LINE SYSTEMS—INTERNATIONAL STUDIES

## Background Information

International organisations such as the CCITT and CEPT are actively engaged in studies concerning the control of errors in digital networks and the error-performance specification of individual digital equipment and systems. The main objective is to define a set of error characteristics for digital equipment, thus ensuring that, regardless of the length and complexity of a digital connection, the overall performance is acceptable in most cases[8].

The article by McLintock and Kearsey in an earlier issue of this *Journal*[4] details the network performance objectives for the ISDN, and these objectives take into account all errors liable to occur. Equipment design objectives for digital transmission systems have to be compatible with the network performance objectives in CCITT Recommendation G821[9]. The network performance objectives in this Recommendation have been formulated as representing the balance between a level of performance satisfactory for most services and the level of performance achievable by using practical transmission systems.

CCITT/CEPT studies have not yet advanced to the stage where the clauses of the Recommendations which are to relate to error performance can be completed. For the present, designers must progress with the development of their systems and, in the absence of complete international guidance, make their own assumptions about the required quality.

In applying the network performance objectives given in CCITT Recommendation G821 to the design of digital equipment, it is necessary to consider the three following factors:

(*a*) The equipment design objective needs to be expressed in a manner which is meaningful and usable to a system designer. Recommendation G821 expresses the overall network objectives in terms of the percentage of error-free seconds and 1 minute having a bit error ratio better than a threshold value; and this is of little value to the designer.

(*b*) There needs to be a margin to allow for errors caused by unknown hazards outside the control of the designer (for example, unquantifiable interference from external sources). These factors would normally be related to the physical and electrical environment in which the equipment was working in and, consequently, their significance would vary from one location to another[8, 10].

(*c*) Due account needs to be given to network planning aspects, the control and responsibility of which are normally with the operating administration.

## Definition of Performance Characteristics

Within CCITT, there is no clear idea as to the best way of formulating design objectives for transmission systems operating over metallic pairs. Although several possible approaches are under study, no single method seems satisfactory in all respects. Two approaches to the error-performance specification have been identified:

(*a*) The error performance of a hypothetical reference digital section (HRDS) of defined length is specified. This serves as a guide to the design of all types of transmission system. This approach is analogous with that adopted for specifying the noise performance of analogue transmission systems.

(*b*) The quality of regenerators used in a transmission system is specified and measured in terms of a regenerator quality parameter.

Although based on different philosophies, it is not to be assumed that either approach necessarily excludes the other. In fact, it looks as if the final outcome of the CCITT studies may result in a methodology in which the two approaches complement each other. This appears to be the most attractive solution, as both approaches are considered essential for a satisfactory performance specification.

### *Specification in Relation to a Hypothetical Reference Digital Section*

Studies are showing that a significant proportion of errors

in digital networks are attributable to effects over which the system designer has little control. Unquantifiable forms of interference (often impulsive in nature) and other so-called *network effects* are dependent, in the main, on the physical and electrical environment, the control and responsibility of which is with the operating administration. Although studies are progressing to improve knowledge of these effects, it will be many years before all the error-inducing interfering mechanisms can be fully quantified and some degree of immunity against such effects provided.

Moreover, it is envisaged that, in some instances, it will not be practicable or cost effective for administrations either to provide total immunity in the system design or to eliminate universally the interference at source. For these reasons, equipment design objectives should ensure a degree of provision in the system design to minimise the effects of such external forms of interference. Therefore, a design objective should logically include a defined conventional and reproducible environment agreed to be reasonably representative of what might be encountered in practice. This approach assumes international standardisation of a conventional environment, a situation which is unlikely to be obtained in practice.

However, until such time as the important error-causing factors can be taken into account in the definition of the environmental conditions forming part of the equipment design objectives, systems need to allow generous margins in respect of them.

Bearing those points in mind, consideration can be given to one possible method of specifying an equipment design objective[8, 11]. Initially, those errors attributable to unquantifiable interference over which the designer has limited control are discounted. It has been proposed, for the purpose of illustration, that 80% of the total permissible impairment be allowed to accommodate these network effects. It seems not possible, as yet, to make firm judgements as to the appropriate margin necessary for a particular system based on current experience. It is convenient to refer to this as an 80% *network discounting factor*. Accordingly, the remaining 20% accommodates the permissible impairments due to the basic system design limitations such as:

(*a*) thermal noise in the case of coaxial digital line systems, and

(*b*) the combined effects of crosstalk from signals operating on other pairs in the same cable for digital line systems that operate on symmetric pair cable.

Such forms of interference are generally considered to be random processes and, consequently, the induced errors, before error extension effects, tend to occur in accordance with a Poisson distribution of intervals.

As an illustrative example (see Table 2), the equipment design objective is derived by using an 80% network discounting factor for a transmission system falling into the so-called *high-grade* circuit classification[4]. The design objective expressed in this way is not particularly useful. Consequently, it is necessary to re-express these objectives in a more meaningful way, so that they can be used by designers. Designers of metallic-pair systems invariably start with the long-term mean error ratio (LTMER), a parameter which has a direct bearing on the noise immunity required at the decision point in a regeneration-type circuit. Given this information, regenerators are designed taking account of the known impairments such as the diminishing of eye opening, noise, crosstalk, level inaccuracies, etc. On the basis that these quantifiable errors are attributable to random processes, the objectives are satisfactorily expressed as an LTMER by using the Poisson model. This manipulation process reveals the equivalent LTMER that simultaneously satisfies the aforementioned objectives (see Table 3).

**TABLE 2**
**Allocation of Percentage of 1-Minute and Errored-Second Requirements for Network Performance and Equipment Design Objectives for Transmission Systems in the High-grade Classification**

Objectives at 64 kbit/s for a high-grade circuit of length 280 km assuming an 80% network discounting factor

| | Percentage of 1-minute intervals with a bit error ratio worse than $1 \times 10^{-6}$ | Percentage of seconds with one or more errors |
|---|---|---|
| Network performance objectives | 0·0448 | 0·03584 |
| Equipment design objectives | 0·00896 | 0·007168 |

**TABLE 3**
**Equipment Design Objectives for a High-Grade Circuit of Length 280 km and having an 80% Network Discounting Factor**

| | Percentage of 1-minute intervals with a bit error ratio worse than $1 \times 10^{-6}$ (*a*) | Percentage of seconds with one or more errors (*b*) | LTMER required to satisfy (*a*) and (*b*) simultaneously at 64kbit/s |
|---|---|---|---|
| Equipment design objectives | 0·00896 | 0·007168 | $1 \cdot 12 \times 10^{-9}$ |

### *Specification in Relation to a Regenerator Quality Parameter*

Specifying the performance characteristics of transmission systems in terms of a line-regenerator-quality parameter has been a traditional method adopted by many national administrations. As mentioned in the earlier section on noise, it has been agreed within the CCITT to specify a crosstalk noise figure for symmetric-pair-based systems and a noise margin for coaxial-based systems[5]. These parameters essentially qualify the regenerator immunity to a dominant form of predictable noise (for example, thermal noise, crosstalk) in addition to incorporating an allowance made by the designer to provide for such effects as temperature variations, component ageing, input signal level variations, etc. The parameter is not effective in indicating the regenerator's immunity against unquantifiable external interference and, therefore, is not an alternative to margin against such interference. These regenerator-quality measures, although useful criteria, are deficient in that they are not easily related to the overall network-performance objective. Regenerator design is but one of the factors contributing to system immunity to error-causing interference. Others need to be brought within the scope of the design objective. Of course, it will be the designer's responsibility to adhere to sound design practices in respect of screening, effective earthing, reducing imbalances on symmetric pairs, high- and low-frequency cut-off in the amplifier etc. in their equipment designs. These measures provide some degree of protection against unwanted external voltages. It can, therefore, be appreciated that the ability of a system to function in the presence of external interference becomes an important attribute of digital equipment. The extent to which a system performs acceptably in an unfavourable environment becomes a principal criterion for assessing more generally the quality.

### Network Planning Aspects

So far, consideration has been given only to design factors *per se*. In practice, it is necessary to take account of the network planning aspects as these often have a significant influence on the actual performance[12]. This is particularly so in the case of digital line systems operating over symmetric pair cables, as the operational performance is dependent on many interrelated factors under the direct control of the operating administration. The factors impacting on such systems are shown in Fig. 2. Consequently, for such systems, it may be sufficient to specify a regenerator-quality parameter (for example, crosstalk noise figure) and not to specify, in addition, a system design objective expressed in terms of LTMER. If such an approach proves to be acceptable, compliance with the network objective will be the sole responsibility of the operating administration.

### Comments

Internationally, it is recognised that the performance specification of transmission systems is far from straightforward. Generally, digital systems, free of faults and in an interference free environment, operate virtually error free. However, in operational circumstances, the extent to which a transmission system performs is largely dependent on the level and significance of the electrical environment. Consequently, it is not possible to define such digital equipment by using a precise methodology as is the case with most equipment (for example, digital multiplex, PCM multiplex etc.). To mitigate against this difficulty, the specification approach formulated by international standard-setting organisations may need to be on two levels:

(*a*) First, the provision of a precise equipment design objective expressed by using the following parameters:

(*i*) specification of performance in relation to HRDS in which the electrical environment is precisely defined, and

(*ii*) specification of a regenerator quality parameter.

(*b*) Second, the provision of explicit guidelines enabling equipment designers to provide adequate immunity against the so-called *network effects*.

Finally, as a cautionary note, it follows that when the performance of an installed system is measured with a view to checking whether a design objective is met, precautions are necessary to identify errors attributable to network effects not covered by the specification.

## ERROR MEASUREMENTS ON PRIMARY-ORDER LINE SYSTEMS

### Background Information

In BT's network, 2048 kbit/s digital line systems operating over symmetric pair cables are used extensively for short-distance applications[13]. The results presented here represent measurements on 14 traffic-carrying systems, monitored continuously for periods of between 3 and 11 weeks, giving a total of over 14 700 system-hours. As there were no criteria for choosing particular systems, there is a good mix in terms of design, route length (between approximately 10 to 28 km), cable type and cable fill, time of year, physical environment, traffic loading, etc.

Notwithstanding this, no statement can be made at this point as to whether these results are typically representative.

Most telecommunication administrations throughout the world have embarked on similar performance surveys with the following objectives:

(*a*) to ensure that existing transmission systems perform to a level of quality consistent with the evolving international error-performance standards for an ISDN, and

(*b*) to characterise the error performance of the evolving digital network in terms that are useful and meaningful to different types of users.

Historically, the assessment of error performance has involved the measurement of the LTMER, a parameter describing the occurrence of errors in a given observation interval. The inadequacy of a single parameter to characterise the error performance is discussed in an article in an earlier issue of this *Journal*[4].

Recognising this inadequacy, the CCITT has recommended (in Recommendation G821) that errors are counted over observation intervals of 1 second and 1 minute, and this has resulted in the concept of error-free second (EFS) and short-term mean error ratio (MER). The results presented in this article conform to the integration periods used in Recommendation G821.

Recording the error data in this manner facilitates the expression of performance in terms of the percentage of observation intervals in which the MER exceeds a threshold value, and this is consistent with the approach adopted internationally. The error performance was determined by counting HDB3 line-code violations during the appropriate intervals. The occurrence of a violation does not define precisely the true error ratio, but, in practice, the rate of occurrence of line-code violations is generally close to the error ratio. Therefore, the terms *error ratio* and *EFS* are used loosely; the results are actually in terms of line-code violations rather than true errors.

The systems were monitored by using an in-house error-logging apparatus, and errors from all sources have been included in the results. Errors falling into the unavailable time have been discounted in accordance with the method agreed internationally[9]. This method states that a period of unavailable time begins when the MER in each second is worse than $1 \times 10^{-3}$ for a period of 10 consecutive seconds. These 10 seconds are considered to be unavailable time. The period of unavailable time terminates when the MER in each second is better than $1 \times 10^{-3}$ for a period of 10 consecutive seconds, and these 10 seconds are considered to be available time.

### Results

Figs. 4 and 5 show the distribution of LTMER and errored

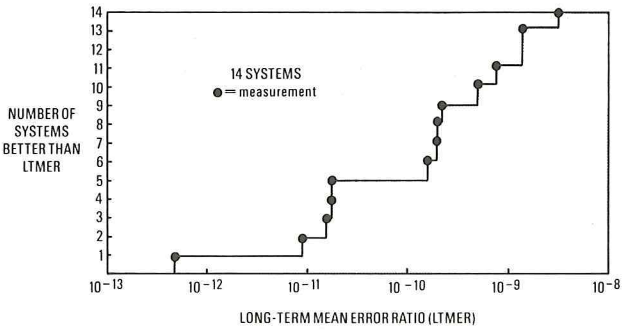



FIG. 4—Error ratio distribution for 2048 kbit/s digital line systems

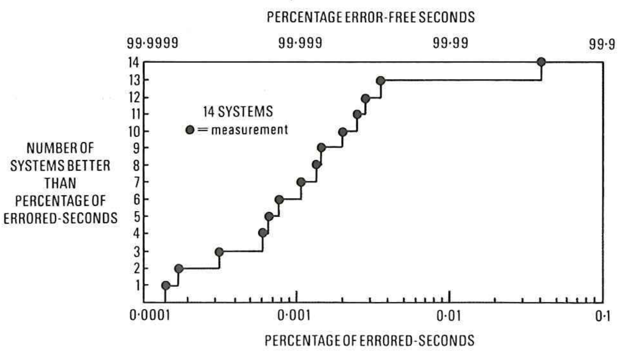



FIG. 5—Errored-second distribution for 2048 kbit/s digital line systems

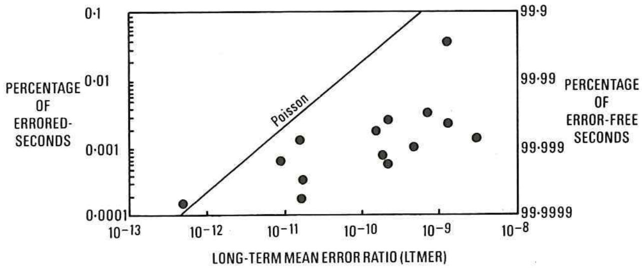
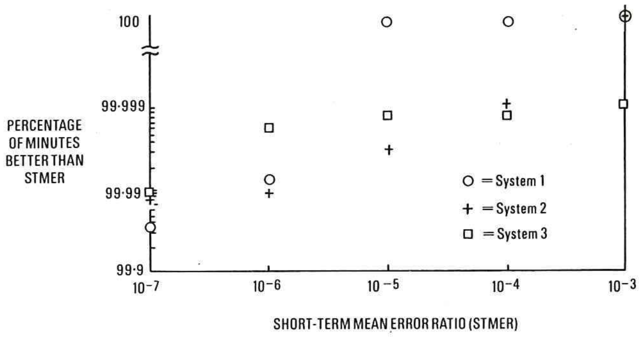



FIG. 6—Scatter plot of seconds with errors against error ratio for 2048 kbit/s digital line systems




*Note*: The short-term MER has been assessed over 1-minute intervals

FIG. 7—Distribution of short-term MER for three selected 2048 kbit/s digital line systems

seconds for all 14 systems. This distribution illustrates the number of systems that had an error performance equal to or less than a certain value of LTMER or errored seconds. The LTMER was determined over the entire observation interval, which varied between approximately 3 and 11 weeks.

Fig. 6 is a scatter plot of the percentage errored seconds versus LTMER. The solid line on the plot represents the result which would be obtained for a Poisson model. The results illustrate the difficulty of establishing a single mathematical model which can be considered representative of all systems. This result mirrors closely the behaviour observed on two hundred and twenty-eight 1544 kbit/s digital line systems surveyed for a total of over 54 000 system-hours in the United States[14, 17].

Fig. 7 shows a selection of distributions of the short-term MER (assessed over 1-minute intervals) for three out of the 14 systems.

Figs. 4 to 7 give a summary view of the survey. Figs. 8

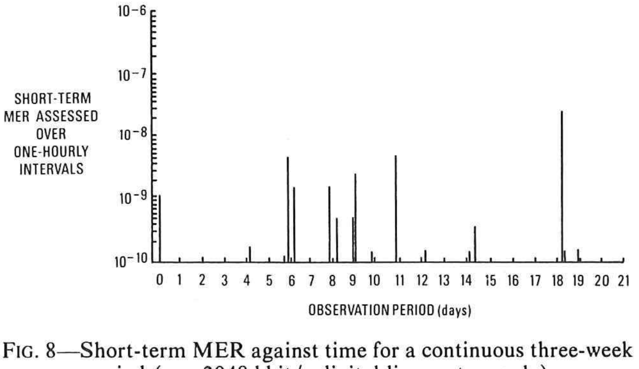



FIG. 8—Short-term MER against time for a continuous three-week period (one 2048 kbit/s digital line system only)

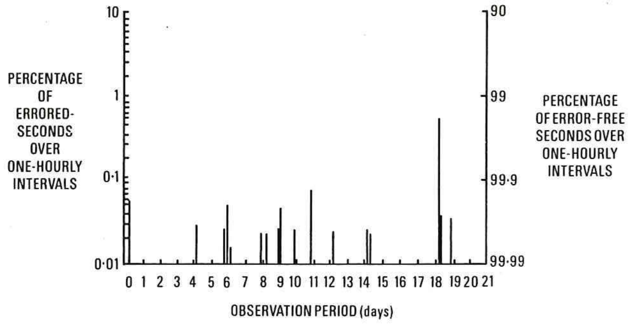



FIG. 9—Percentage of errored-seconds against time for a continuous three-week period (one 2048 kbit/s digital line system only)

and 9 give a more detailed view of one particular system over an arbitrary 3-week period. They show the distribution of hourly MER and hourly percentage errored seconds with respect to real time. These diagrams illustrate the point that for most of the time a fault-free system (that is, a system complying with the design specification) is virtually error free.

### Comments

Based on the survey to date, the results are encouraging in that the operational performance is judged to be good and consistent with the objectives established internationally. A more detailed inspection of the raw data indicates varying error characteristics and it seems clear that several competing error-inducing processes cause this widely time-variant performance on each system. On certain systems, the error characteristics are predominantly random in nature and exhibit very few error bursts. It is believed that such systems are limited by near-end crosstalk, and that the physical environment is not severe or that the system is immune to the external interference presence. On the other hand, certain systems are characterised by the frequent occurrence of error bursts with discernible background errors due to near-end crosstalk. Again it is difficult to judge whether this behaviour results from the inability of the system to cope with unquantifiable interference from unknown sources or whether there was insufficient margin against impairments assumed in the design.

The major deficiencies in using LTMER as a quality measure were exemplified. An apparently unacceptable LTMER was due, in the main, to infrequent transient effects causing short-duration error bursts containing many errors. In reality, the quality was much better than implied by the LTMER and, in fact, for a significant proportion of the time, the system was error free, as shown in Fig. 5.

## ERROR MEASUREMENTS ON HIGH-SPEED LINE SYSTEMS

### Background Information

BT has ninety-eight 120 Mbit/s digital line systems in the trunk network operating over 1·2/4·4 mm coaxial pair cable. These systems were designed to use the existing 12 MHz coaxial cable routes and have the same repeater spacing of approximately 2 km. The results presented here are for seven 2048 kbit/s digital paths (including associated multiplex) routed over three 120 Mbit/s digital line systems. These paths were monitored continuously for periods of between 2 and 10 weeks, giving a total of 4591 system-hours. Once again there was no criteria for picking a particular system and there is a good mix in terms of route length (66, 85 and 260 km), time of year, physical environment, traffic loading etc. The results presented are for a very small sample, and no statement can be made as to whether these results are typically representative for this type of

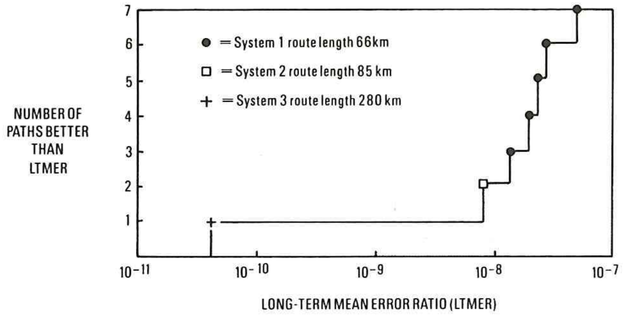
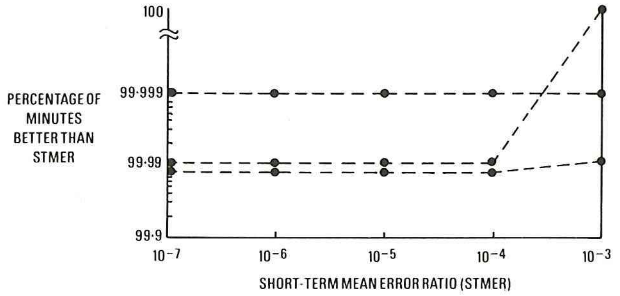



FIG. 10—Error ratio distribution for 2048 kbit/s digital path routed via a 120 Mbit/s digital line section




*Note*: The short-term MER has been assessed over 1-minute intervals

FIG. 13—Distribution of short-term MER for three selected digital paths

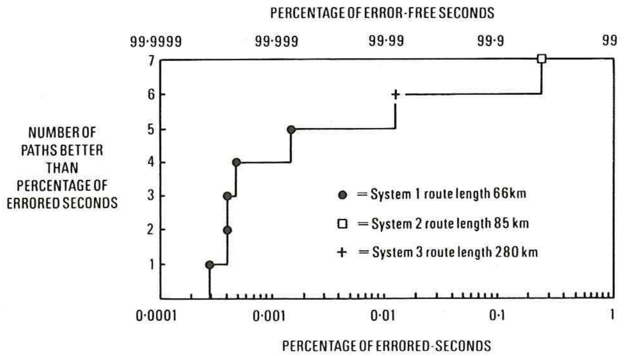



FIG. 11—Errored-second distribution for 2048 kbit/s digital path routed via a 120 Mbit/s digital line section

network configuration. Measurements were made on non-traffic carrying 2048 kbit/s digital paths, looped at the far end, using a $2^{15}-1$ pseudo-random binary sequence. Error detection was based on a bit-by-bit comparison. Errors from all sources have been included in the results, with the exception of those falling into unavailable time which have been discounted in accordance with the method agreed internationally[9].

### Results

Figs. 10 and 11 show the distribution of LTMER and errored seconds for the seven 2048 kbit/s digital paths. This distribution illustrates the number of paths that had an error performance equal to or less than a certain value of LTMER or errored seconds. The LTMER has been derived over the whole measurement period, which varied from 2 to 10 weeks.

Fig. 12 is a scatter plot of the percentage of errored seconds versus LTMER. The solid line represents the results which would be obtained for a Poisson model.

Fig. 13 shows the distribution of short-term MER (assessed over 1-minute intervals) for three of the seven 2048 kbit/s digital paths.

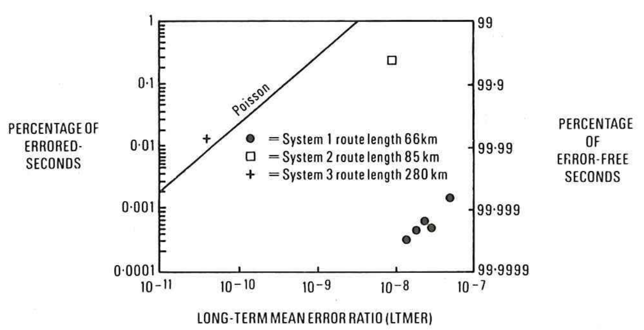



FIG. 12—Scatter plot of seconds with errors against error ratio for a 2048 kbit/s digital path routed via a 120 Mbit/s digital line section

### Comments

A detailed examination of the raw data indicates that most errors are concentrated within relatively short time intervals. This intense short-term error activity is believed to arise through the so-called network effects, and results in a performance which is not necessarily length dependent. The intrinsic quality of the equipment appears to be satisfactory and of a high standard, with the internal random error-inducing processes (for example, thermal noise) making only a small contribution towards the overall error performance. At this juncture, it is not possible to isolate the equipment within the digital path (that is, digital multiplex or digital line systems) that contribute most of the errors. The error performance does not necessarily represent the performance of a 120 Mbit/s line system *per se*.

As with the results presented for primary-order line systems, the use of the LTMER as a performance indicator can give a misleading picture of the performance. Investigations are continuing in an attempt to isolate network effects that cause errors.

## ERROR RESULTS BY OTHER ADMINISTRATIONS

Within the last few years, administrations throughout the world have initiated programmes to monitor the performance of field-trial and operational digital transmission systems. This has mainly been the result of a growing realisation among CCITT members of the difficulty of recommending error-performance parameters without considering the real performance of digital systems in the field. It has not been possible in this article to survey the results that other administrations have published, but a list of references has been included[15–25] for the interested reader. While these references are by no means exhaustive, they should allow the reader to gain an appreciation of the performance of other administrations' digital networks.

## ERROR MEASUREMENTS FOR THE DIGITAL LOCAL AREA

### Background

The move towards the digital local area (DLA) aims at providing digital transmission right into the customers' premises. A bit rate of 80 kbit/s in each direction is sufficient to provide more than the basic telephone service, but the desire is to use the DLA to provide wider bandwidth services. It is the error performance of the inexpensive metallic pairs

used in the local area that prevents easy application of such services, especially visual services. The problem is that more and better simultaneous facilities require a higher bit rate, leading to higher bandwidth and greater signal attenuation, and this is aggravated by the need for simultaneous bothway transmission on the same pair. The alternative of installing dependent repeaters within subscriber loops is expensive and operationally difficult. The importance of achieving satisfactory digital transmission on the subscriber loop has been recognised by successive International Symposia on Subscriber Loops and Services[26].

Subscriber pairs have been found to give satisfactory interference immunity for traditional voice-frequency use with more economical installation practices than those used in the inter-exchange network[27]. Impedance discontinuities are found at the frequent changes of pair type within a subscriber loop and at bridged taps in some countries, and abnormalities such as rectified split pairs are commonly encountered. Other interference that couples into the significant unbalance of such a path comes from sources such as other similar customers, carrier systems, radio transmissions and impulsive interference from switching and signalling and even electric power lines. The severity of impulsive noise on digital transmission in the local network has long been recognised[28]. To achieve the error performance allocation for the local loop in the presence of such interference is a major challenge if the number of repeaters is to be minimised.

It would be too expensive and disruptive to customers to make comprehensive measurements of these time-varying effects on all conditions of local loop plant. Statistical surveys have been attempted, for example, by American Telegraph and Telephone Company[29], but the pattern varies widely between countries and even between local areas within any one country. Consequently, an increasing number of DLA transmission trials are also being mounted. Although these trials cannot have the scale to replace the statistical surveys of interference, they do give information on the acceptability for DLA use of a modest number of local loops.

### Early Results

At this early stage of DLA application, few results are available and these correspond to a wide variety of experimental techniques to achieve bothway transmission.

The trials that are most advanced are those based on the simple, but generally uneconomic, arrangement of separate pairs for the two directions of transmission. Some results were published as early as 1978[28]. Perhaps the most extensive trial has been the provision of digital subsets on 56 lines at Horsens, in Denmark, in a project[30] carried out by the International Telegraph and Telephone Corporation and Jutland Telephone Company. Most lines were between 0·5 km and 2 km in length, with a maximum of 3·5 km. The average attenuation of the 80 kbit/s signal was about 10 dB. Each line was monitored for at least one full working day, and virtually all lines were found to have very low LTMER and to achieve better than 99·9% error-free seconds, thus meeting the likely network requirement. The main source of consistent errors appeared to be impulse noise disturbances. A small proportion of the lines were observed over much longer periods.

After this trial in Jutland, the same equipment was used in more extensive measurements in a Strowger PABX environment at Standard Telecommunication Laboratories. These results, as yet unpublished, confirmed that impulsive noise caused by conventional telephony signalling on other cable pairs was the prime cause of errors, and that, although the levels of interference varied with the exchange type, under adverse conditions, reasonable network requirements could be met. With this sort of interference, it becomes appropriate to include the relevant telephony traffic calculation in the digital cable-fill planning calculations.

With regard to bothway transmission on the same pair, the most common arrangement is *burst mode* or *time separation*, which uses an increased clock rate to shuttle blocks of bits alternately in each direction[31]. This technique has been investigated in the USA on unloaded loops[32], in Japan[33], in West Berlin[34], and in many other countries. However, such reports generally do not cover a sufficient time period with unselected subscriber loops for statistically significant conclusions to be drawn of performance. The other popularly supported arrangement is the echo-cancelling hybrid, using adaptive techniques in which the same frequency band is usable for bothway transmission simultaneously[31]. Because of the advanced nature of this approach, field performance results are limited.

In the UK, interest has been shown in digital subscriber carrier systems, whereby a conventional voice-frequency circuit is retained on the same pair used for bothway digital transmission. The results of very limited trials in the BT network are encouraging, but do not allow statistical conclusions to be drawn[35].

### Comments

The limited DLA transmission measurements to date show that errors tend to occur in bursts with long error-free periods in between. These tests suggest that many unloaded subscriber loops can be worked bothway at around 80 kbit/s with acceptable error performance without independent repeaters. However, the results for one country or area cannot be applied to another with different subscriber loops practices, so that extensive trials are needed in each case. Although impulsive noise features strongly as a prime limitation, radio-frequency interference, inter-system crosstalk and pair discontinuities are also real problems and must be considered in the design. Minimisation of radio-frequency interference problems to and from DLA systems can be achieved only by strict planning of the spectrum. For satisfactory error performance to become routine, planning rules must be established by the particular administration to ensure good transmission and noise conditions. These rules should take account of impulsive noise by using the traffic statistics applicable to the particular cable.

## CONCLUSIONS

Digital line systems operating over symmetric pairs are required to show high immunity to near-end crosstalk from similar systems in order that repeater spacing and cable fill may be maximised: although near-end crosstalk does commonly impose a limitation in practice, results presented here and elsewhere show that there are other sources of errors causing systems to have a time dependent performance. Higher-rate systems operating over coaxial pairs are usually designed to combat thermal noise in order to maximise repeater spacing: the results presented for 120 Mbit/s show that most of the errors are concentrated within relatively short time intervals. First experiments with DLA have also resulted in burst errors that tend to be traceable to man-made interference; in particular, signalling on conventional voice-frequency circuits in the same cable.

In all cases, there are serious deficiencies in the traditional use of LTMER as a quality measure. A system with an apparently unacceptable LTMER due to infrequent transient effects may well work error free for the vast majority of time and so be quite acceptable. This is now accepted within CCITT, which is recommending appropriate quantitative limits. There has to be a distinction between the overall requirements on a system to meet the network objectives, inclusive of unknown sources of interference, and the more stringent requirements regarding the known sources against which the design is made. Administrations should be more active in the difficult task of collecting field information on error causes, and provide the necessary data to enable system

designers to provide the appropriate immunity. It would be desirable if standardised tests for external interference could be devised.

## ACKNOWLEDGEMENTS

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## Biographies

Richard Catchpole graduated from Cambridge University in 1965; in 1966, he received a Master's degree from the Massachusetts Institute of Technology. After three years at the research laboratories of English Electric Computers, he joined Standard Telecommunications Laboratories, where he has worked for the past 14 years on digital line transmission. He has instigated a variety of techniques for regenerative repeaters involving equalisation, the 6B-4T code and fault location. For the past three years he has been a department manager responsible for landline systems. He is now manager responsible for the transmission systems division.

Peter Norman received a B.Sc. in physics at Birmingham University in 1948. From 1943 to 1945 he worked at the Signals Research and Development Establishment on direction-finding equipment and, in 1946 and 1947, did similar work for the Plessey Company. He then joined the Transmission Division of Standard Telephones and Cables, where he worked on the development of landline systems, including open-wire, coaxial, PCM and subscriber carrier systems. More recently he has been concerned with optical-fibre systems.

Stephen McElvanney graduated from Edinburgh University in 1981 with an honours degree in mathematical physics. Later that year he joined British Telecom, where he has been involved in work on digital transmission network standards and, specifically, the error performance of digital networks.

# Service Standards for Packet-Switched Networks—An Introduction

P. R. H. BURRINGTON, B.SC., M.SC., A.M.I.E.E.†

UDC 621.394.4 : 681.327.8 : 389.64

*Grade of service is a familiar concept in the field of circuit switching. Its definition and measurement for packet-switched networks is currently a subject for discussion within the CCITT. This article outlines the relationship between network architecture and grade of service for such networks.*

## INTRODUCTION

Telecommunications networks exist to meet the needs of the customers who use them. While tariffs should reflect the value placed on the service provided, they should also generate a reasonable return on capital invested. It would not be impossible to provide a network whose quality, in terms of performance and reliability, would satisfy all of its customers all of the time; a group of competent engineers could probably do it, but at considerable cost.

However, every marginal improvement in quality has to be paid for by a corresponding increase in capital invested, and it follows that this extra investment will generate a proper financial return only if the customer considers the increased benefit to be worth the tariff that accompanies it. If this is not the case, he will take his business elsewhere.

To make an accurate estimate of the point at which marginal increases in investment cease to pay for themselves, business skills of a high level are required. This difficult task is made almost impossible when the return on investment includes social or other intangible benefits to the business and to the community as a whole.

Nevertheless, an estimate is made of the requirements for performance and reliability which together form the heart of the network specification. It is usually expressed in terms of the probability that a customer will find the network in a state such that the quality of service is unacceptable in some respect.

At this stage, the engineer is presented with the specification and asked to design a network that will please most of the people most of the time. The methods of accomplishing this by using circuit-switched technology are well established. The same cannot be said, however, of packet-switched networks where the concept of quality embraces a far wider spectrum of interdependent customer-perceived variables, which are affected by network architecture and loading in ways that are often difficult to analyse.

Nevertheless, the CCITT* has studied the problem of establishing realistic standards of quality and has submitted Recommendations covering transit delay and blocking for final approval by the October 1984 CCITT Plenary Assembly. The following account describes, in outline, how British Telecom's contribution to these studies was formulated, and how the planner might expect to use the resources at his disposal in order to meet those Recommendations.

## GRADE OF SERVICE

Network dimensioning and design generally consist of using the minimum resources that will satisfy the majority of customers. Customers' requirements may be many and varied, but usually fall into one of two broad categories: those satisfied by the functional characteristics of the network (that is, what it does), and those satisfied by the consistency with which the characteristics are offered (that is, how far it can be relied upon to do it).

For instance, a telephone network may offer direct dialling to a particular country. If a customer finds that an unacceptably high number of his call attempts fail, then the service may be considered satisfactory in the first sense, but not in the second.

Consistency is expressed in terms of quality of service (QOS); this indicates the likelihood that network functions will be unavailable for use by the customer, either because of congestion resulting from an overload, or because of equipment failure. In this sense, available means able to provide a specified minimum level of performance. Unavailability, therefore, includes occasions when network functions are available only in an unacceptably degraded form.

A subdivision of QOS is grade of service (GOS), which specifically indicates unavailability owing to shortage of network resources under congestion conditions. A GOS specification for a given traffic load is, therefore, the starting point in the network dimensioning part of the design process.

## CIRCUIT SWITCHING VERSUS PACKET SWITCHING

In the circuit-switched environment found in most existing telephony networks, all resources required for a call are allotted at the time that the call is set up, the customer retaining exclusive use of them for the duration of the call. The most important of these is transmission bandwidth, and any contention for it that results from an increase in demand takes place when the connection is requested. Although the transmission quality of the allotted channel may subsequently be degraded by fluctuations in load, its entire bandwidth remains at the disposal of the customer for as long as he cares to pay for it, and regardless of the extent to which he uses it.

Packet-switching technology allocates resources to each individual packet as and when the packet requires them. Each packet of customer's data has to compete with others for bandwidth and switch processing time as it makes its way across the network. Queues are provided to accomodate those packets waiting for service; indeed, queueing is an intrinsic feature of packet switching. The bursty nature of data traffic is deliberately exploited to effect economies in overall transmission capacity. This characteristic produces a peak-to-mean ratio in the load which is much larger than that found in telephony traffic and allows the planner to provide bandwidth sufficient to accomodate little more than the mean level, while relying on queues to cope with the

† International Business Services, British Telecom International

* CCITT—International Telegraph and Telephone Consultative Committee

severe, but short-term, fluctuations that occur. Network design to meet a GOS specification consists largely of engineering these queues to match a given traffic load.

Any packet transmitted during the course of a call is consequently affected by the levels of traffic load that exist at points along its route.

Definitions of GOS can be made specifically for the cases of blocking and delay as follows.

(*a*) *Blocking*

(*i*) For circuit switching: the probability that a request for connection finds no free trunk available and no free place in the queue (if queueing is permitted).

(*ii*) For packet switching: the probability that a *call request* packet finds no free logical channel available or no free place in the queue for a resource; and

the probability that a data packet finds no free place in the queue for a resource.

(*b*) *Delay*

(*i*) For circuit switching: the probability that unacceptable delay occurs between request for connection and connection. (This applies only where queueing is permitted.)

(*ii*) For packet switching: the probability that any individual packet experiences unacceptable delay while in transit through the network.

## QUEUEING THEORY

Queueing delay is a random variable, and can therefore be described only in terms of the likelihood that its value will lie within a given range. The probability distribution function shown in Fig. 1 is of the negative exponential type. It gives the probability that $x$, the time spent queueing, will equal or exceed the value indicated by the x-axis. It is the type applicable to a single-server queue and a Poisson arrival process†.

Two features of the function should be noted. Firstly, it is certain that this queueing period is zero or greater. Secondly, the graph is asymptotic with the x-axis. This indicates a small but finite probability that even the longest queueing delay will be exceeded. The function can be used to show that 95% of the values of $x$ will be less than a value which is approximately three times the mean value of $x$. This is an important result and will be used later.

Given a value of mean queueing delay, the probability that some given value of delay will be exceeded can be calculated. Alternatively, given a probability that a certain delay is exceeded (that is, a specified GOS), the calculation can be carried out in reverse to deduce the mean delay associated with that GOS specification.

All other things remaining equal, the length of a queue will shorten as the time taken to process each member of the queue is reduced. This time is known as the *service time* and can be measured or calculated for a processor switching a packet and for the transmission of a packet by an outgoing port.

The term *load* refers not to the traffic load, but to the proportion of available time that the server spends actually servicing packets. It is usually expressed as a mean value taken over some interval which is large compared to the service time. Instantaneous values will either be zero or one, depending on whether or not the server is doing anything at that instant, and are of little use.

Now that the distribution function has been used to derive the mean delay associated with a particular GOS specification, it only remains to obtain a value of service time.

The relationship between service time, load and mean delay is shown in Fig. 2, and is derived from the Khintchine-Polloczek theorem of single-server queues†.

Thus, given a specification for GOS, the maximum permissible service time for a given load, or the maximum permissible load for equipment with a given service time can be calculated.

GOS specifications have been made by the CCITT with respect to acceptable transit delays for the various packet types. Two values are quoted in each case: the mean delay, and the delay which may be exceeded by no more than 5% of all the packets transmitted. They relate to hypothetical reference connections defined in CCITT Recommendation X92, which are intended to represent typical real connections likely to be found in practice.

Blocking probability emerges naturally from the analysis as the probability that the queue length will become greater than the length of the physical queue provided.

† MARTIN, J. Design of Real-Time Computer Systems. Prentice Hall, 1967.

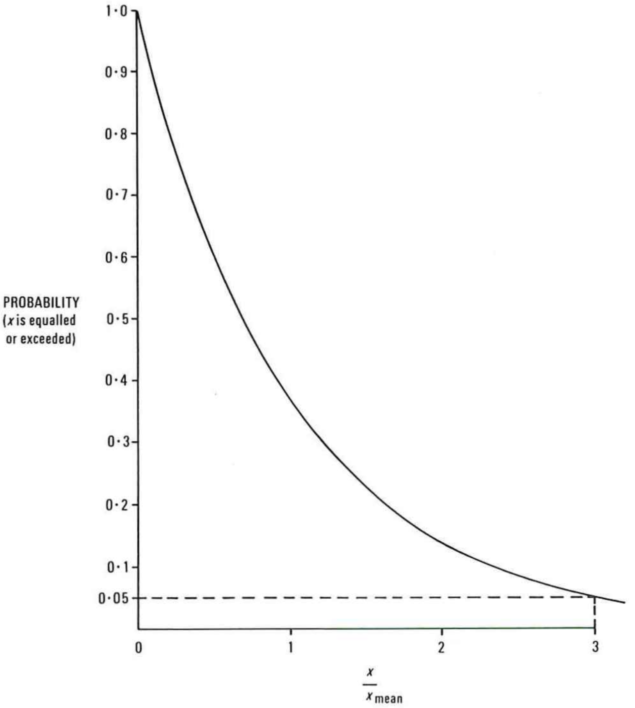



FIG. 1—Probability distribution function

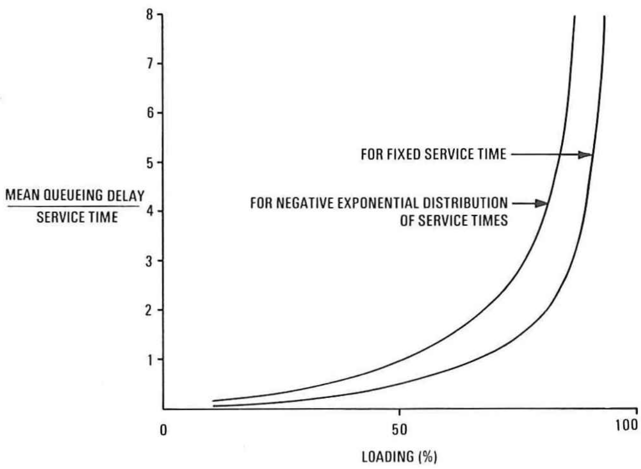



FIG. 2—Relationship between service time, load and mean delay

## SINGLE QUEUES AND SUCCESSIVE QUEUES

Every packet that passes through a switch under load can be expected to encounter a queue for the use of each process required to be carried out on it. At the very least, there will be two of these processes. The first switches the packet to the outgoing port, and the second transmits it. If the switching processor is required not only to switch but also to manage or even to carry out the buffering of the incoming packet, then the number of processes and, hence, queues will increase. A single packet-switching exchange may comprise a number of switches and several exchanges may take part in the connection; so the total number of successive queues along a virtual circuit is likely to be large.

Given the overall mean delay due to queueing for such a circuit, is it reasonable to derive a GOS as if there were an equivalent single queue contributing the same mean delay? It would be extremely convenient if the statistics were identical for the two cases in this respect. However, this is not true, as the following argument will demonstrate.

Consider a series of $N$ queues, each contributing a mean delay of $D/N$ seconds. The overall mean delay is therefore $D$ seconds.

A single queue causing a mean delay of $D$ seconds offers, as shown in Fig. 1, a 0·05 probability that a delay of more than approximately $3 \times D$ seconds will be contributed. (This is equivalent to a GOS specification permitting no more than 5% of packets to experience a queueing delay of more than this time.)

What is the probability that the series of queues having the same mean delay $D$ overall will contribute $3 \times D$ seconds delay? This requires each queue to contribute $3 \times D/N$ seconds, which is three times its mean delay.

The only difference between the probability distributions of long and short queues is in the scale of the x-axis. The ratios between the various statistics such as mean and GOS remain constant. The probability that a particular value of delay exceeds three times the mean value is 0·05, regardless of the size of that mean value.

So the probability that an individual member of the series contributes its $3 \times D/N$ seconds of delay is the same as that for the single queue to contribute $3 \times D$ seconds, namely 0·05.

The probability that every member of the series makes this contribution simultaneously to produce the required total of $3 \times D$ seconds is therefore 0·05 to the $N$th power: a far less likely event than the single queue contributing the same delay.

The correctness of this argument in more general circumstances can be proved mathematically. Probability distribution functions for some values of $N$ are shown in Fig. 3; their derivation is not described in detail here.

It is clear that as the number of queues, $N$, increases, the likelihood of the delay exceeding a given value, such as the GOS value, diminishes. Therefore, if a given overall mean queueing delay can be shared among a larger number of queues, each with a correspondingly lower mean delay, the effect is to improve the GOS from 95% to some higher percentage. Also, a higher overall mean load can be accommodated while still keeping within the 95% GOS specification, although the mean delay increases as a result.

The overall effect of increasing the number of queues is to reduce the spread of probable delay values and with it the level of uncertainty in predicting them. Thus, with departures from the mean becoming less and less likely, the performance of the network with respect to delay can be predicted with greater and greater confidence.

## ECONOMIC CONSIDERATIONS

Service resources, such as processing power and transmission bandwidth, must be combined with queueing resources, in the form of buffer storage (memory), to meet the specification in the most economic way.

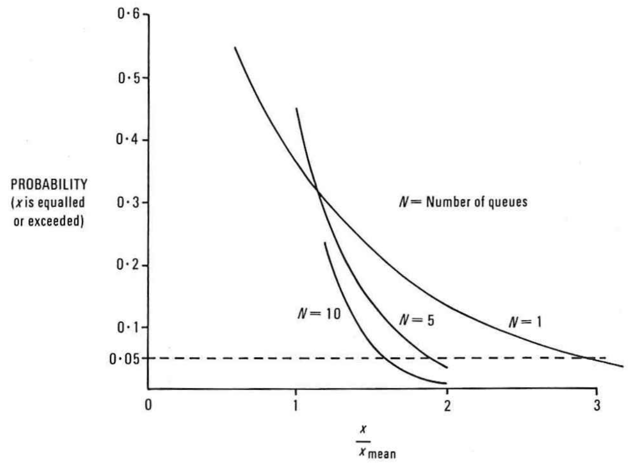



FIG. 3—Probability distribution functions for a series of $N$ queues

Often several satisfactory solutions to the problem are possible, because trade-offs between service and queueing requirements can be made. Queue length, and hence buffer requirement, increase with load, given a particular service time. Traffic load applies a load on the server which is proportional to the service time. This load can be reduced by reducing the service time, a secondary effect being the shortening of the queue.

The shortening of service times requires more or faster processors and greater transmission bandwidths, the increased cost being offset by reduced buffer requirements. Conversely, the economies made by choosing slower transmission and processing equipment may be negated by extra expenditure on buffer capacity to accommodate the resulting longer queues.

GOS restrictions will ultimately place a limit on the long-queue policy. Nevertheless, there is still room for flexibility in design.

Once queueing delays have been estimated, it only remains to add these to the fixed delays, such as service times for switching and transmission, together with propagation delay, to give the total for a particular route.

## CCITT RECOMMENDATIONS

Work carried out within Study Group VII of the CCITT during the period 1981–1984 has established GOS requirements for delay in packet-switched networks, and these are to be approved and published shortly as Recommendation X135. Mean and 95% values of acceptable delay are recommended for *call request*, *call accepted/connected*, *data*, and *clear request* packets over the shortest and longest hypothetical reference connections defined in Recommendation X92.

Maximum permissible blocking probabilities are covered in Recommendation X136.

## PRACTICAL CONSIDERATIONS

The usefulness of the mathematical model as a planning aid depends upon the ease with which its various parameters can be measured in a live network environment.

Service and queueing times for both switching and transmission resources, traffic loading and loading of servers must all be susceptible to measurement. Moreover, most of these parameters are random variables which can only be described mathematically in terms of their statistics. Consequently, many measurements of each variable are needed to obtain this information.

The ideal environment is, therefore, one in which traffic loading is sufficiently heavy to exercise the queueing mechanisms and is also statistically stationary. This latter condition means that, although short-term fluctuations in value are expected, the statistics of those values (such as mean value and probability distribution), are independent of time. Measurements of delay can thus be made in the confident knowledge that any statistics derived from them would be identical with those obtained at any other time. These conditions are rarely found in practice.

Experience with the international packet-switched service (IPSS) network has shown that traffic loads fluctuate considerably with the busy periods throughout the day. They rarely remain stationary for long enough to allow sufficient measurements to be taken. However, sensible interpretation of incomplete results has yielded useful information.

GOS is normally specified with respect to the delay incurred between one customer's interface and another's. However, to measure and record the transit delay of every packet is a formidable undertaking. One method is to time-stamp all packets as they enter the network and to record the elapsed time as packets leave; but to do this, synchronised clocks would be required in all exchanges on the network boundary. In addition, the transmission of measurements to some central point might be sufficient to change the load patterns under observation, as could the extra data carried by each packet. Similar problems are likely to arise if artificially generated test packets are used in sufficient numbers to be effective, although this is an established method on some networks.

Examination of the more specific parameters reveals some further problems. Service time for transmission is simply the packet length divided by the line speed. The model assumes a negative exponential distribution of packet lengths, this being the worst case with respect to delay. In practice, the length and its distribution are likely to depend on the habits and preferences of the customer as well as the nature of the information being transmitted. However, it is reasonable to expect processor service time to be constant for a given packet type, although task scheduling policies within the software may introduce a distribution of values.

Instantaneous queue lengths can be monitored by the switch software, provided that separate physical or logical areas of buffer are provided for each queue. The IPSS goes some way to achieving this by allowing the total instantaneous buffer capacity to be monitored.

Load on a server is probably the easiest parameter to measure, as it merely consists of the proportion of time that the server is active. A processor can readily be monitored by programming it to run a clock during its idle moments, although if it executes tasks other than those directly associated with the switching of packets then this measurement might be misleading. Transmission-path idle time is simply that which occurs between transmission of frames, and is easily measured externally by using commercially-available test equipment.

## CONCLUSION

Although exhaustive measurements have yet to be completed, initial results tend to confirm that the CCITT Recommendation X135 can be achieved in practice and represents a level of quality which is acceptable to users.

### Biography

Paul Burrington graduated from Salford University in 1975 with an honours degree in Physics and joined British Telecom as an Assistant Executive Engineer. Since 1978 he has worked on the Euronet and IPSS packet-switched networks, during which time he graduated from Essex University with an M.Sc. in Telecommunications Systems. He recently participated in the development of CCITT Recommendations on packet-switching grade of service and is currently involved in the provision of a second gateway exchange for IPSS.

# Book Review

*Fiber Optics Communications*. Edited by Henry F. Taylor. Artech House, Inc. xiv + 331pp. £37·00

This book is a collection of reprints taken from the Journals of the learned societies, the IEE and IEEE, and from conference presentations. The volume covers by chapter: General Review, Fiber Fabrication, Propagation in Fibers, Fiber Buffering and Cables, Connectors and Splicing Techniques, Light Sources and Transmitters, Photodetectors and Receivers, Multiplexing Components and Switches, System Related Noise Effects, System Design and Experimental Systems and finally, Field Trials and Operational Systems. Indexing of the material consists solely of a list of contents; there is no detailed subject or author index. The editor's introduction briefly reviews the field historically and traces the development of the major themes presented in the volume.

The contents of the volume indicate that the latest material dates from late 1982, which suggests a copy date for the contents at about the close of 1982. The only exception to this is a group of papers from the IEEE Journal on Selected Areas of Communications dated April 1983. However, the manuscripts of these date from a similar time to the rest since that issue was delayed in publication. Thus the data presented is roughly two-years old; and includes little on the design of 1500 nm wavelength systems.

The selection of papers shows a generally good balance between advances in research and applications, and reflects the general activity in the field, although the editor's selection largely reflects the late awakening of interest in single mode fibre and systems in the USA by comparison with the UK and Japan, since very little of their detailed design or operation is covered.

The volume shares with all reprint or multi-author works the difficulty that, because each contribution stands alone and is disconnected from the whole, a balanced picture becomes very difficult to obtain. The volume is aimed at a competitive market, one in which both the IEE and the IEEE have already published a number of reprint volumes. Against these, it is probably more up to date than the others, apart from the recent IEE volume devoted only to Electronics Letters reprints. Furthermore, for staff within British Telecom (BT), the volumes of reprints published annually in BT by the Optical Communications Division at Martlesham provide further competition. Thus, this volume is undoubtedly worth acquiring for reference purposes in a library, but individuals planning to purchase it are recommended to look at the competition and make the choice that best appeals to their pockets and interests. Each offering is different, partially overlapping and partially complementing, and it is not easy to identify any one volume as being clearly supreme.

J. E. MIDWINTER

# Performance of Digital Radio-Relay Systems

G. HART, B.TECH., C.ENG., M.I.E.E.†

UDC 621.374 : 621.396.4.029.64

*This article outlines some of the different types of digital radio-relay systems that are now in world-wide service, identifies the factors which contribute to their performance, and illustrates the performance standards that are currently being achieved.*

*The CCIR* performance standards for high-capacity digital radio systems are discussed, together with some of the system planning guide-lines in current use, and these are compared with world-wide results from a number of system field evaluations. The comparison shows that the CCIR standards can be achieved on established microwave networks by using advanced equipment designs which incorporate diversity and adaptive equalisers to overcome the effects of dispersive fading.*

*The performance and availability of medium- and low-capacity systems are also briefly reviewed, and predictions of non-availability due to rainfall fading are given for some typical systems.*

*The article concludes that the performance standards for digital radio systems will be in accordance with the highest quality grades required for transmission media in an integrated services digital network.*

*This article is a revised version of an article that first appeared in The Radio and Electronic Engineer‡, and is reproduced here by permission of the Institution of Electronic and Radio Engineers. This updated version takes account of the Interim Meeting of CCIR Study Group 9 held in May 1984.*

## GENERAL BACKGROUND

Most radio-relay systems in the world at present operate in the frequency bands below about 10 GHz and use analogue frequency-modulation methods. Many countries have now established extensive microwave networks, and the radio media is used to support a wide variety of different transmission services including telephony, with capacities of up to 2700 telephony channels per radio bearer, and television. The high standards of performance and availability achieved on these systems are now generally taken for granted, and are founded on more than 30 years of operating experience. It is against this general background that digital radio is now being introduced by many telecommunication authorities throughout the world.

It is not obvious that digital transmission is a good choice for systems such as radio relay that are confined to operate in restricted bandwidths. Digital modulation does not always result in the most efficient use of available spectrum, nor in the most economic equipment realisations. The problem of spectrum efficiency is exacerbated by the relatively inefficient pulse-code modulation coding technique at present adopted for telephony, which uses 64 kbit/s per voice channel, and therefore can be very wasteful of radio-frequency (RF) bandwidth. The most economic radio transmission solution, particularly for long-haul systems, stems from maximising the capacity of each radio bearer; until recently, an analogue format, using single-sideband amplitude modulation to give bearer capacities up to 6000 telephony channels, was considered by some Administrations to be the optimum economic realisation[1].

Analogue systems, however, suffer from the problem of accumulation of noise and other impairments at successive repeaters, and this factor can significantly limit the performance of long-haul systems, or systems in higher-frequency bands, which use much shorter hop lengths. One of the prime technical advantages of digital radio is the ability to regenerate at every repeater, and this overcomes many of the problems of accumulated impairments.

Perhaps the greatest interest in digital radio was stimulated by the world-wide trend towards all-digital networks, where the benefits of an integrated network, with digital switching and signalling, far outweigh any of the disadvantages of using digital radio. In Japan, the USA, and Europe, intensive programmes of research and development have been in progress since around the early-1970s. Evolution has gradually taken place in system design, and this has capitalised on new RF technology to introduce more complex modulation methods which overcome the spectrum efficiency problems, and on new component technology to permit higher capacity working.

Many different types of system configuration, capacities and modulation methods, are now offered as proprietary designs; these have resulted from the many different priorities and requirements set by those who have made first use of digital radio. Systems are available for use in the local, regional, trunk, and international sectors of the network.

Different digital hierarchical structures have been adopted for use in North America, Japan, and Europe, as shown in Table 1, and this has also lead to diversification.

In the USA, trunk radio systems with the non-hierarchical capacity of 2 × 45 Mbit/s are common. This is, in part, because of the spectrum management regulations imposed by the Federal Communications Commission (FCC), which have tended to concentrate attention on system designs that permit band sharing with existing analogue plant on the

**TABLE 1**
**Digital Multiplexing Hierarchies**

| Hierarchical Level | Bit Rate (Mbit/s) | | |
|---|---|---|---|
| | Europe | North America | Japan |
| 1 | 2·048 | 1·544 | 1·544 |
| 2 | 8·448 | 6·312 | 6·312 |
| 3 | 34·368 | 44·736 | 32·064 |
| 4 | 139·264 | 274·176 | 97·728 |
| 5 | | | 397·200 |

† Trunk Services, British Telecom National Networks
* CCIR—International Radio Consultative Committee
‡ HART, G. The performance of digital radio relay systems. *The Radio and Electronic Eng.*, Apr. 1984, **54**(4), pp. 155–162.

same route, rather than those designs that offer maximised band capacity[2]. In Europe, 2 × 34 and 140 Mbit/s capacities are more common for trunk systems, but these capacities are not always easy to accomodate within the frequency channeling arrangements for analogue systems recommended by the CCIR. Several new channel plans, specifically optimised for totally digital use, are therefore currently under consideration by some Administrations.

In the early stages, little was known about the behaviour of digital modulation in anomalous propagation conditions, and a great deal of elaborate field testing has been carried out on a world-wide basis to examine the performance capabilities of these various system configurations. This has lead to a much better understanding of the behaviour of digital systems, and to some important refinements in order to make performance more robust, particularly during multipath propagation. Most importantly, it has enabled a significant amount of data on the performance of digital radio systems to be collected, some of which is reported in this article.

## TYPES OF SYSTEMS AND APPLICATIONS

### High-Capacity Systems

High-capacity digital radios for trunk network applications mainly use the 4, lower-6, upper-6, 7, 8, and 11 GHz frequency bands, where rainfall attenuation is not too significant, and hop lengths in excess of 50 km can often be used. European trunk route lengths are around 150–200 km, made up from hops of about 40–50 km average length. Fig. 1 shows the British Telecom (BT) microwave network, which is fairly representative of trunk radio networks in Europe.

In North America, a distinction is made in the performance requirements for short-haul trunks, generally up to about 400 km length, and long-haul trunks, which include the East–West transcontinental routes up to 6500 km.[3] Fig. 2 illustrates the structure of the common carrier trunk network in the USA at 6 GHz.

Depending upon the frequency band, modulation methods and channel capacity used, a full system could comprise up to 12 bothway channels, usually arranged in an $N+1$ protection switched configuration. It is normal practice to provide regeneration at each repeater.

Linear orthogonally-polarised emissions are normally used for radio-relay systems, so that interference between adjacent channels can be controlled by the cross-polar discrimin-

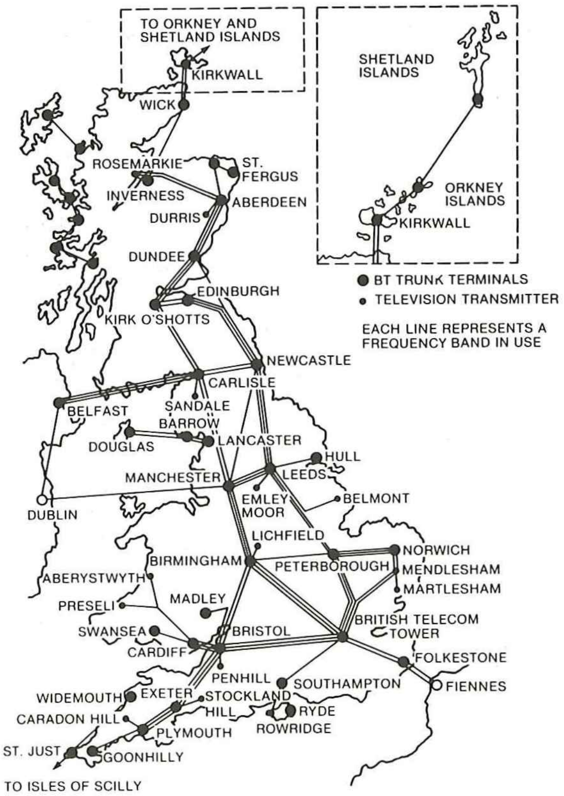



FIG. 1—The British Telecom trunk microwave network

ation (XPD) of the antenna system. Some types of system, optimised for high spectrum efficiency, use two channels on the same frequency assignment (Fig. 3), and are thus heavily dependent upon XPD to ensure good standards of performance[4].

A summary illustrating the range of different types of system in service is given in Table 2.

The main performance impairment for these trunk systems comes from the effects of anomalous propagation, and interference from band sharing with other systems, such as the fixed satellite service. Multipath fading is by far

FIG. 2—Common carrier trunk network in the USA (6 GHz)

## TABLE 2
## Different Types of Trunk Digital Radio in Service

| Channel Capacity | Frequency Band | Modulation Method | Maximum Number of Bothway Channels | Source | Date in Service |
|---|---|---|---|---|---|
| 45 Mbit/s | 4 GHz | 8PSK | — | USA | 1980 |
| 2 × 32 Mbit/s | 11 GHz | 8PSK | — | Japan | 1981 |
| 2 × 34 Mbit/s | 4 GHz | 8PSK | 6 | Europe | 1980 |
| | L6 GHz | 8PSK | 8 | Europe | 1983 |
| | U6 GHz | 8PSK | 8 | Europe | 1983 |
| | 7 GHz | 8PSK | — | Europe | 1983 |
| | 11 GHz | 8PSK | 12 | Europe | 1983 |
| 2 × 45 Mbit/s | 6 GHz | 16QAM | — | USA | 1981 |
| | 8 GHz | QPRS | 6 | Canada | 1980 |
| | 11 GHz | 8PSK | — | USA | 1980 |
| | 11 GHz | 16QAM | — | USA | 1981 |
| 3 × 45 Mbit/s | 6 GHz | 64QAM | — | USA | 1984 |
| 140 Mbit/s | 4 GHz | 16QAM | 6 | Europe | 1985 |
| | 4 and L6 GHz | RBQPSK | 8 | Europe | 1984 |
| | U6 GHz | 16QAM | 8 | Europe Australia | 1984 |
| | 11 GHz | QPSK | 6 | Europe | 1982 |
| | | 8PSK | 8 | Europe | 1983 |
| | | 16QAM | 12 | Europe | 1985 |
| 200 Mbit/s | 4 GHz | 16QAM | 7 | Japan | 1984 |
| 400 Mbit/s | 19 GHz | QPSK | — | Japan | 1979 |

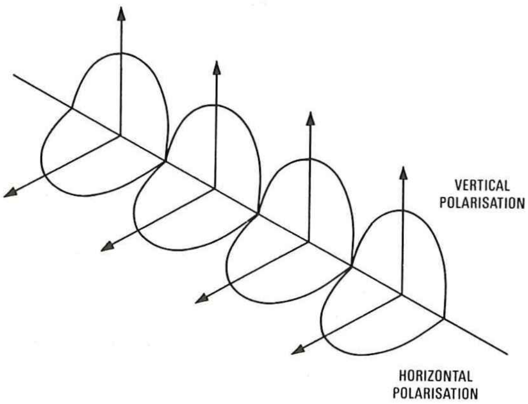



FIG. 3—Co-frequency cross-polar channel arrangements

the most significant problem, and the large fade margins required to give protection from the effects of multipath normally result in very high standards of performance in normal propagation conditions, which exist for most of the time (typically more than 99% of a year).

### Medium-Capacity Systems

Digital radio systems with capacities of less than about 45 Mbit/s are now in common world-wide use for regional transmission requirements. In Europe, these systems are used for spur and junction connections, and typical applications comprise one or two hops, about 25–30 km in length, using 1 + 1 channel configurations[5]. An example of this type of use is shown in Fig. 4, which illustrates the French inter-city microwave network[6]. Medium-capacity digital radio is also now being used in metropolitan areas to provide media diversification, and to cater for topographically difficult transmission requirements[7].

The frequency bands above about 12 GHz are ideally suited to short-route applications, since hop lengths become restricted by rain attenuation and depolarisation above this frequency. Sufficiently long hop lengths can be obtained above 13 GHz for multipath still to be a problem that can affect performance, but, at 18 GHz and above, practical hop lengths are less than 15 km, and atmospheric multipath is not very important. For systems affected by rainfall, non-availability, therefore, becomes the important design criterion, and performance objectives are usually easy to meet.

### Local Network Applications

Until now, digital radio has not found much use in local networks, mainly because of the comparatively high radio equipment costs. Recent innovations in the development of point-to-multipoint systems[8], and point-to-point systems, however, have shown that radio can be effective in this application, by providing a versatile alternative where the more traditional transmission methods are not easily made available, and, importantly, by providing short lead times to the provision of service.

Such systems are now finding world-wide use in both city and rural environments, catering for data and telephony requirements. Point-to-point systems with 2 and 8 Mbit/s

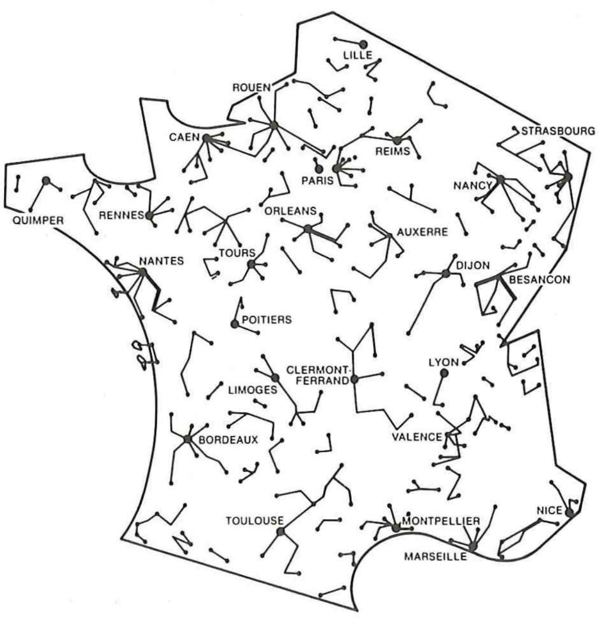



FIG. 4—The French inter-city microwave network

capacities in the 18 GHz band are in common use in the UK, operating between roof-tops of prominent buildings, over path lengths of up to 15 km. Time-division multiple access (TDMA) point-to-multipoint systems will also shortly be introduced in city environments, over links of up to 10 km, to cater for the specialised service requirements of the business community.

Systems designed for local networks are usually very simple, in order to reduce costs and increase reliability. Many operate in the higher-frequency bands and are, therefore, subject to rainfall attenuation problems, rather than multipath. Protection switching is usually not used so that non-availability from both rainfall attenuation and equipment unreliability is the principal design factor.

## ANOMALOUS PROPAGATION AND PERFORMANCE

There are several different forms of anomalous propagation which can affect the performance of digital radio systems[9], but multipath and precipitation fading are normally the dominant causes of performance degradation. In CCIR and CCITT† terminology, the distinction is made between performance and non-availability by using an event-duration criterion of 10 s. For this reason, multipath propagation mainly affects transmission quality, whereas rain fading limits availability.

### The Incidence of Multipath Propagation

Multipath is an unavoidable propagation problem for any radio system which uses hops of more than about 20 km. When multipath conditions occur, propagation paths with differing delays are combined together at the receiving antenna and result in large amplitude and phase distortions in the transmission channel. Multipath is often accompanied by significant mean reductions in the received signal level (>20 dB) because of effects such as defocusing and abnormal refraction.

The meteorological conditions which bring about multipath are normally seasonal and relatively infrequent, occurring mainly during night-time hours in the hot summer months, when windless conditions allow air at different temperatures to form into atmospheric layers, resulting in refractive-index anomalies.

Monthly statistics are used to express multipath propagation data, because this period normally contains a high proportion of the total yearly activity. The concept of a worst month is used in order to take account of the yearly statistical variations, and the definition of 'worst month' has been given by the CCIR in Recommendation 581. Typically, a worst month would comprise four or five nights with activity, and the cumulative total activity for the year would be about three times that of the worst month. Fig. 5 illustrates the yearly distribution of error activity caused by multipath on an 11 GHz 140 Mbit/s radio system[10].

The incidence of multipath fading during the worst month has been found empirically to be proportional to about the fourth-power of the hop length[9], and is greater on hops which span flat regular terrain.

Several hops within a digital radio section may experience multipath conditions at the same time, but it is statistically unlikely that errors on any radio channel result from simultaneous multipath events on different hops. For this reason, the performance of a digital section is normally found by summing the periods of error activity of each hop, rather than summing error ratios.

### The Effects of Multipath Propagation

The differences in delay between these multiple propagation

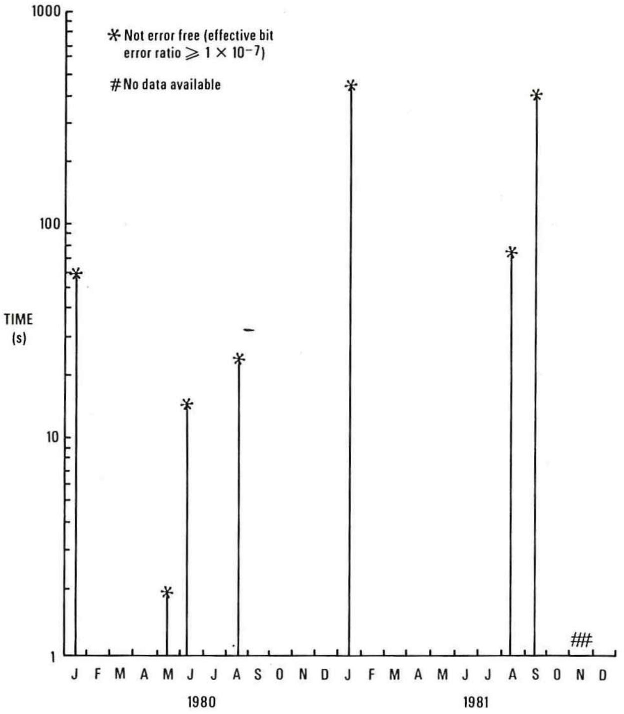



FIG. 5—Yearly distributions of error inducing multipath propagation taken from an 11 GHz 140 Mbit/s system evaluation in the UK

paths are also a function of hop length and antenna beamwidth, ranging between about 0·5 and 6 ns, which is significant in comparison to the pulse duration of high-capacity systems. The amplitude and phase distortions appearing in the transmission channel as a result of a simple two-path situation are illustrated in Fig. 6, although much more complex situations are also possible. The attenuation and group delay notch responses shown in Fig. 6 also appear at other harmonically related frequencies and, if the delay is sufficiently long, they may be repeated elsewhere in the band, and therefore cause other channels to be affected simultaneously.

These amplitude and phase distortions bring about intersymbol interference, and any mean depression fading also present enhances the noise in the receiver, prior to regeneration.

Interferences from adjacent and cross-polar channels, and from other systems, may also increase significantly, because the cross-polar discrimination of the antenna system also reduces in proportion to fade depth, and because fading on the wanted signal does not correlate with that on interfering paths from other systems. This reduction in antenna cross-polar discrimination is particularly significant in contributing to the performance degradation of systems which utilise co-frequency cross-polar working[11] (two channels on the same frequency assignment).

Errors in transmission are therefore caused by a wide range of different combinations of the three basic impairments of noise, interference and intersymbol interference, that may occur during fading conditions.

The impairments are unstable and vary rapidly, since any small atmospheric perturbations may significantly affect the notch characteristics, or move it to a different frequency (channel). For this reason, error events coming from multipath fading are usually of short duration, and primarily affect performance.

A characteristic common to most digital radio systems is the rapid transition from virtually error-free performance to loss of alignment, as the threshold conditions for these impairments are exceeded during a fading event. The resulting degradation is known as *outage* (usually defined

† CCITT—International Telegraph and Telephone Consultative Committee

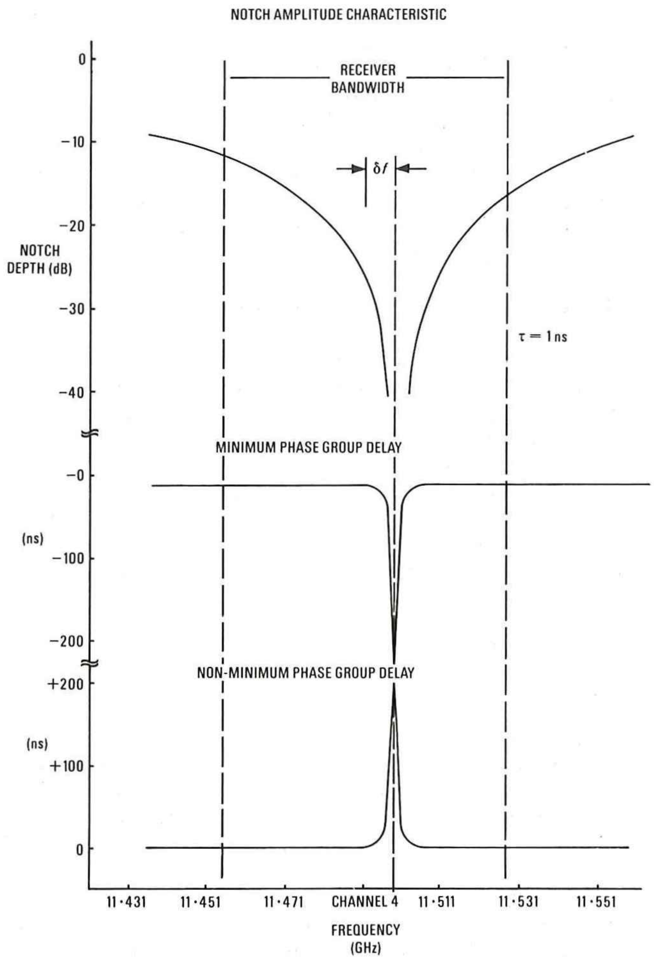
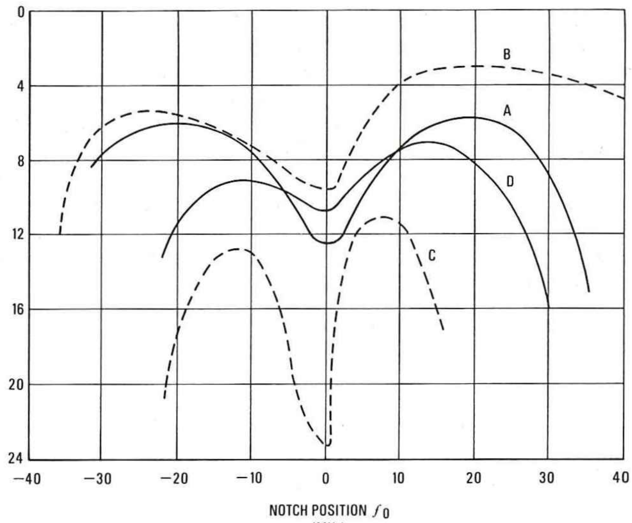



FIG. 6—Two-path fading amplitude and delay distortions




Curve A: 78 Mbit/s, 8 PSK ($\alpha$ = 0·3), 26 Mbaud
B: 90 Mbit/s, 8 PSK ($\alpha$ = 0·5), 30 Mbaud
C: Same as B but with amplitude slope and notch adaptive equalisers
D: 90 Mbit/s, 16 QAM ($\alpha$ = 0·5), 22·5 Mbaud
$\alpha$: Roll-off factor
$\tau$ = 6·3 ns
Bit error ratio = $10^{-3}$

FIG. 7—Signatures for digital radio-relay systems

as bit error ratio (BER) $> 1 \times 10^{-3}$ for more than 1 s), and typical outage events on any channel may last for up to about 10 s[10, 12].

### Equipment Behaviour in Multipath Conditions

Sensitivity to multipath impairments can be considerably different for different system designs, depending principally upon capacity, choice of modulation method, and channelling arrangements adopted. Many early radio system designs failed to take sufficient account of the effects of multipath signal distortions and were found during field evaluation to have unacceptable standards of performance[13]. Since that time, much work has been taking place to develop adaptive signal-processing techniques which are effective in correcting for these distortions, thus giving more robust performance during fading conditions.

The sensitivity of a system design can be evaluated in the laboratory by using the technique known as *signatures*[14], in which a family of curves is obtained describing the locus of simulated multipath conditions which result in outage. Some typical system signatures are shown in Fig. 7. These sensitivity curves are useful for comparing the merits of different equipment, and can be used to obtain performance predictions, if sufficiently detailed and comprehensive propagation data are readily available[18].

### Rain Fading

Rainfall attenuation increases significantly with frequency to the extent that it is necessary to reduce hop length above about 12 GHz in order to provide adequate fading margins. For this reason, the bands above 12 GHz tend to be used for regional and local network applications, rather than for overlaying existing trunk routes.

Large fade margins are required to protect against the heavy rainfall intensities which occur intermittently, even in the temperate latitudes such as the UK. The worst rainfall conditions occur in cellular distributions up to about 15 km diameter mainly during thunderstorm activity[15], and the more frequent uniform rainfall distributions (for example, warm-front conditions in the UK) have a much less significant effect.

Prediction methods have been developed to enable the attenuation effects of rainfall to be calculated on a cumulative annual basis[16], so that, for a given hop length and system fade margin, it is possible to calculate the average yearly non-availability. Table 3 shows a comparison of the non-availability due to rainfall of several UK systems currently in service.

## PERFORMANCE OBJECTIVES

### CCIR Recommendation 594

The work so far completed by CCIR Study Group 9 mainly addresses high-capacity digital radio systems operating above the second hierarchical level. Recommendation 594, concerning allowable BERs, and Recommendation 557, concerning availability, have so far been approved. These Recommendations refer to a 2500 km hypothetical reference digital path (HRDP), given in CCIR Recommendation 556, which was chosen (in accordance with CCITT Recommendation G104) to be the most appropriate network model for describing trunk radio applications, and followed similar

**TABLE 3**
**Comparison of Non-Availability Caused by Rainfall Attenuation**

| System | Hop Length | Typical Fade Margin | Non-Availability Bothways per Annum (%) |
|---|---|---|---|
| 13 GHz 34 Mbit/s | 30 km | 40 dB | 0·001 |
| 18 GHz 8 Mbit/s | 10 km | 30 dB | 0·001 |
| 28 GHz 8 Mbit/s | 10 km | 30 dB | 0·01 |

precedents set for analogue radio performance Recommendations. This 2500 km entity is also used as the basis for expressing performance impairments coming from band sharing with the fixed satellite service[17].

Recommendation 594 was revised at the May 1984 Interim Meeting of CCIR Study Group 9[27], mainly to take account of the evolution of work in the CCITT on Recommendation G821 (see below). The new provisional limits in Recommendation 594 now include both error-ratio and errored-second objectives. These are specified for a unidirectional 64 kbit/s channel as follows:

(*a*) error ratio should not exceed $1 \times 10^{-6}$ for more than 0·4% of any month measured with a one minute integration time,

(*b*) error ratio should not exceed $1 \times 10^{-3}$ for more than 0·054% of any month measured with a one second integration time, and

(*c*) errored seconds should not exceed 0·32% of any month.

As far as their effect on system planning is concerned, these new objectives are not significantly different from previous versions and, while presently only provisional, they are likely to be adopted in full at the XVIth Plenary Assembly of CCIR.

The numerical values of clauses (*a*) and (*b*) were derived from two considerations. Firstly, the outage criterion, given in (*b*), resulted from a consensus of world opinion concerning the performance standards that could be achieved on existing networks at realistic economic levels. In the BT 11 GHz network, it has been found necessary to implement space diversity on about 60% of hops to achieve this standard[10], and a more stringent requirement could, in some cases, introduce prohibitive costs; for example, if new intermediate stations were necessary to shorten hop lengths.

Secondly, account was taken, as far as possible, of the work completed by CCITT Study Group XVIII on the performance requirements for the integrated services digital network (ISDN), given in CCITT Recommendation G821. Revisions to Recommendation G821 carried out in the May 1984 meeting of CCITT Study Group XVII[19] now incorporate high-, medium- and local-quality grades of circuit, at appropriate dispositions within the overall 27 500 km hypothetical reference circuit (HRX), and the overall performance criteria have been subdivided to give network requirements for each quality grade. It is now evident that CCIR Recommendation 594 is consistent with the highest-quality grade requirement of the HRX.

Some North American Administrations have established alternative performance objectives, which are expressed as non-availability requirements from all causes, including multipath, rain-fades, equipment unreliability (the terminology is different to that adopted by the CCIR). For example, the TransCanada Telephone System (now Telecom Canada) reference circuit, consisting of 144 hops and 16 switching sections, is 6560 km long, and the total non-availability allowance on this circuit for one year is 103 minutes[20].

For the present, the CCIR recommendations do not cover radio systems used in the medium- and local-grade sectors of the ISDN, but a new report[27] has recently been prepared by CCIR Study Group 9 on the performance and availability standards of systems intended for local network use. Network error-performance objectives for local- and medium-grade circuits are included in CCITT Recommendation G821.

### Trunk Network Planning Objectives

Many Administrations already use Recommendation 594 as the basis for deriving planning objectives for high-capacity digital radio systems. In the absence of any proper guidance as yet from the CCIR, it has become an accepted practice to subdivide time rather than error ratio, so that a digital radio section of length $L$ km would have an outage allowance of $0{\cdot}05L/2500\%$ of any month.

Because of the highly non-linear variation in outage with hop length (see later in this article in the section on measurements of multipath outage), this objective is strictly applicable only to entire routes, and not to individual hops. In practice, some advantage can be gained from off-setting differences in performance where a mixture of hop lengths is involved.

The process of system planning on an established network is iterative. Firstly, the performance of each hop is estimated, taking into account the characteristics of the basic system design, and a number of hop parameters, including length, path profile, geographical location, and all known sources of interference. Secondly, and successively, performance-improving techniques such as diversity and adaptive equalisation are added to the worst hops, until the route outage criterion is met. Because of the difficulty in accurately predicting outage, a more conservative approach is often adopted by using simple practical rules for the use of these facilities[10].

## BAND SHARING WITH THE FIXED SATELLITE SERVICE

The frequency bands used for trunk radio systems are also heavily used by the fixed satellite services, and this creates problems of mutual interference. Satellites use the bands in pairs: one for up-links (for example, L6 GHz), and one for down-links (for example, 4 GHz). Trunk radio systems, however, use the bands bi-directionally, and this gives rise to two main sources of interference, as shown in Fig. 8. Fortunately, a well organised set of co-ordination rules has been established in the Radio Regulations (Appendicies 28 and 29), which serve to control the interference to mutually acceptable levels.

There is usually a constant level of interference into terrestrial system receivers coming from the aggregate of many off-beam, but visible, interfering transmitters (geostationary satellites). For practical purposes, it is usually assumed that the combined effect of these multiple sources is equivalent to thermal noise. At the levels of power flux density (PFD) given in CCIR Recommendation 358-3, the effect of this interference is to make a marginal reduction in the fade margin of each hop, and hence to increase outage during fading conditions.

During some anomolous propagation conditions which occur for only very small periods of time (ducting etc.), other transmitters (ground stations) which are normally

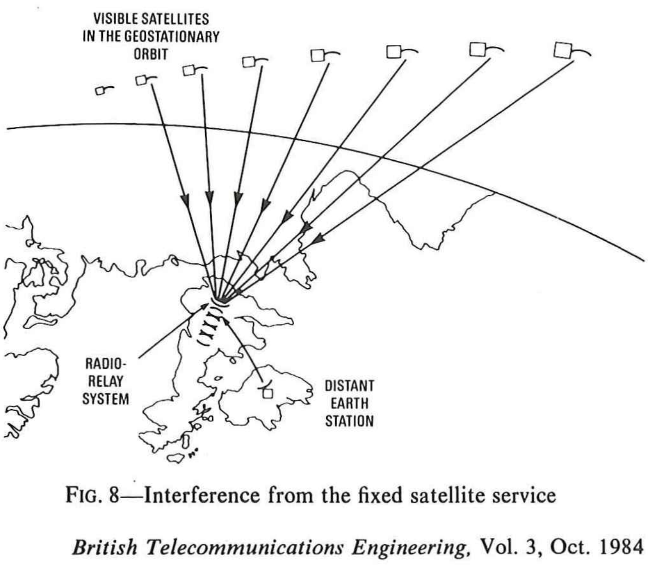



FIG. 8—Interference from the fixed satellite service

obscured by the horizon may be received in beam and create quite high levels of interference. For the earth station equivalent isotropic radiated power (EIRP) given in CCIR Report 386-3, this interference by itself may be sufficient to cause degradations in performance. The problem is not common to all radio-relay receivers, and only two exposures are usually assumed to occur in a 2500 km HRDP.

Allowances, therefore, have to be built into calculations of trunk system performance to take account of these band-sharing problems. For the present, the CCIR envisages that about 10% of radio systems outage will result from this cause[17].

## MEASUREMENT PERFORMANCE RESULTS

### Sources of Data

A considerable amount of published material on radio performance is now available, but mainly in the form of observations made of single-channel radios operating over experimental hops, chosen for their high incidence of fading activity. It is quite difficult to extract representative data from these results, or to undertake any meaningful analysis, because of the differences in measurement methods, observation periods, and presentation. Much of this work was purely investigation into the effects of propagation on the behaviour of equipment, and was not specifically intended to produce performance data in accordance with the criteria of established recommendations.

Some of the first experimental work on high-capacity systems was carried out in the USA by Bell; the performances of several different radio designs were compared on a, by now famous, 42·5 km hop in Palmetto, Georgia [13, 14]. Other performance assessments have been undertaken in France (140 Mbit/s 8PSK[21]), the UK (140 Mbit/s QPSK and RBQPSK[10, 22]), Japan (200 Mbit/s 16QAM[23]) and Italy (140 Mbit/s 16QAM[12]).

Performance results from a 90 Mbit/s 365 km digital radio section in Canada, and a 90 Mbit/s 220 km radio section in the USA, were published in References 18 and 22 but, in general, very little information is available on the measured performance of operational systems over complete digital radio sections.

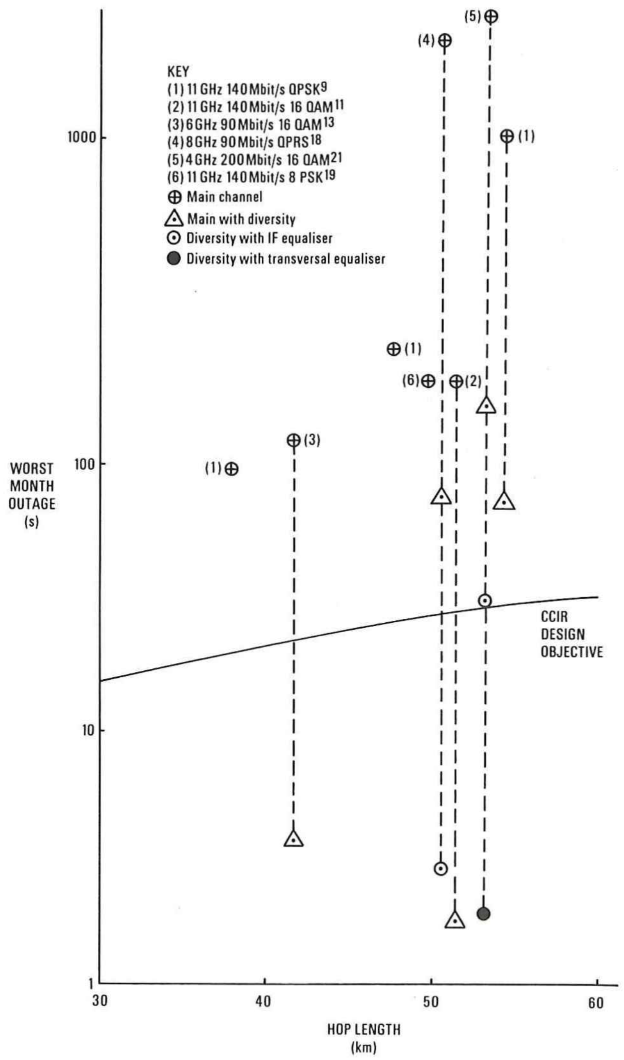



FIG. 9—Worst-month outage for high-capacity digital radio

### Measurements of Multipath Outage

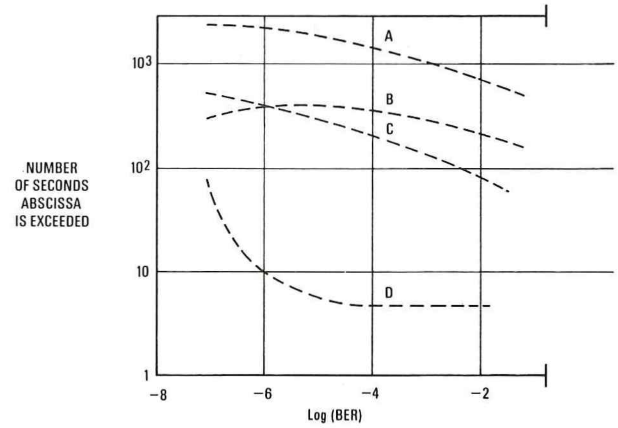
Fig. 9 compares some of the measured worst-month outages on single hops from the published material. The results shown are taken from a variety of different hops (including some over-water paths); therefore, it is difficult to compare the merits of different systems from this figure. The outage planning criterion discussed in the section on performance objectives is also shown for comparison.

Two important observations can be made from these results. First, over the range of hop lengths most commonly used for trunk systems (30 to 60 km approximately), most systems have some difficulty in meeting the CCIR objective without the use of performance-improving techniques such as space diversity and adaptive equalisers (see later in the section on performance-improving techniques).

Second, performance in terms of the worst-month outage worsens in a highly non-linear fashion with increasing hop length. In practice, therefore, the outage of a digital radio section is dominated by one or two bad hops; in general, those which are the longest.

### Error Ratio Distributions

Fig. 10, taken from Reference 25, illustrates the type of error ratio distribution commonly exhibited by digital radio systems. The Figure shows that the period of time for which an error ratio of $1 \times 10^{-6}$ (Clause 3.1 of CCIR Recommendation 594) is exceeded is less than 10 times that of the period of outage; this is a fairly common characteristic of high-capacity digital radio systems.




Frequency: 6 GHz
Bit rate: 78 Mbit/s
Modulation: 8PSK (roll-off factor $\alpha = 0{\cdot}5$)
Channel spacing: 30 MHz
Hop length: 42·5 km

Curves A: Unprotected hop 3 m (10 ft) horn reflector antenna
B: Space diversity with IF combiner
C: Adaptive IF amplitude slope equaliser
D: With B and C

Total observation period = 38 days (August–September 1978)

FIG. 10—Example of error ratio distributions for high-capacity digtial radio systems

The outage criterion of the CCIR objective is, therefore, usually the most stringent to meet during system planning, and other objectives are normally easily fulfilled once this has been met.

For convenience, error ratios are normally measured at the system bit rate, not at 64 kbit/s, and widely different integration periods have generally been used for BER measurements. Because of the influence of demultiplexing, however, the errored-second performance cannot be measured directly at the system bit rate, and practical assessments of this parameter are therefore quite complex.

### Long-Term Background Errors

As indicated previously in this article, digital radio systems are designed to operate normally more than 20–30 dB above the threshold conditions, and the long-term background error ratio performance, in common with other media, is therefore mainly dependent upon the non-deterministic environmental effects, such as extraneous sources of impulsive interferences (lightning discharges, power supply surges etc.).

The long-term error ratio for the 365 km 90 Mbit/s systems discussed in Reference 20 was about $1 \times 10^{-13}$, measured over a 4-month period and excluding the effects of propagation. The average error-free interval over this period was 16·29 hours.

Very few results have been published on the error-free-second performance parameter.

### S Curves

An alternative and simpler method of predicting system performance to that mentioned earlier in the section on anomalous propagation and performance comes from the concept of an effective fade margin.

Fig. 11 shows the behaviour of a system measured over a long period under representative field conditions, and gives the probability that multipath events, described only by reference to fade depth, result in outage. Fig. 10 shows how the probability of outage increases over a range of fade depths from about 20–35 dB as the properties of each multipath event become more critical. A probability of outage of 100% is reached at a fade depth much less than that corresponding to the thermal noise margin of the hop. This S curve function has been found to be relatively independent of hop characteristics.

The S curve thus gives the effective fade margin of the system. Together with readily available propagation data on the probability of occurrence of fading during the worst month, it can be used to give a prediction of system worst-month outage on most hops[26].

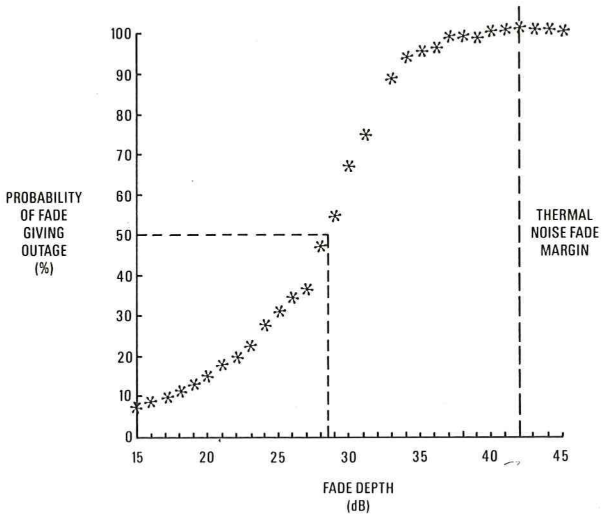



FIG. 11—Concept of effective fade margin

## PERFORMANCE IMPROVING TECHNIQUES

### Use of Space Diversity

Space diversity involves the use of an additional receive antenna at a different height to the main antenna, as shown in Fig. 12. In multipath conditions, the direct path lengths to the receive antennas are approximately equal, but the indirect path lengths differ by a small amount; optimum diversity spacing occurs when this difference is an odd number of half wavelengths. By combining the signals from each antenna in phase, or by switching between antennas, the period of time for which fading is seen by the receiver can be reduced.

The diversity improvement factor has been assessed in the field for several systems, and some of the results are shown in Fig. 9. In general, diversity can reduce outage time in the worst month by a factor of the order of 30 times, but the degree of improvement depends upon both system and path parameters.

Space diversity, however, is generally used selectively, for it can be an expensive technique both in terms of equipment costs, and in terms of providing the accommodation on the tower for any additional antennas required.

### Use of Adaptive Equalisers

Several different approaches to the design of an adaptive equaliser are currently available; some are more effective than others. In general, each type is tailored to a specific system design.

Frequency-domain techniques attempt to correct mainly for amplitude slopes or notches, by processing the intermediate-frequency (IF) signal. This form of equalisation is, therefore, not ideal, since it fails to fully-correct for phase distortions. However, it is cheap to realise and, in practice, can easily be added retrospectively to most system designs. Relatively small improvements in outage, of the order of 5–10 times, have been reported when these IF techniques are used in isolation but, significantly, a much bigger total improvement (as high as 800 times on some systems) has been found when they are used together with space diversity.

Baseband equalisers take the form of decision feedback demodulators or transversal equalisation, which attempts to correct for intersymbol interference both from preceding and following pulse patterns. These baseband techniques are relatively new and complex, and are still under investigation. Some significant improvement factors have been measured on 200 Mbit/s systems in Japan[23].

### Use of Multiline Protection Switching

The $N + 1$ protection switching systems used on trunk radio provide the facility of frequency diversity as an added bonus. If the channel experiencing a fading event can be made to switch transparently to a protection channel before the error ratio deteriorates excessively, then some improvement in

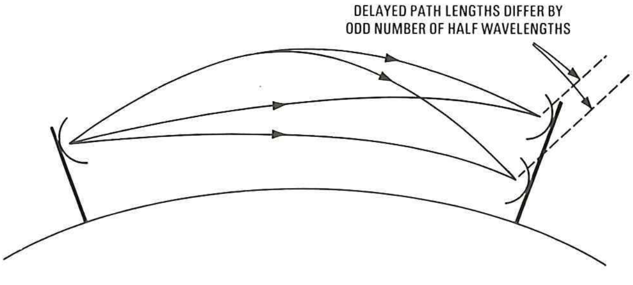



FIG. 12—Space diversity in multipath propagation conditions

performance can be achieved. Protection switching systems for trunk radio are therefore normally designed to be hitless; that is, facilities are provided for fast assessment of channel error ratio by means of parity bit insertion, and for bit sequence alignment prior to switch change-over.

Improvements in performance of the order of six times have been reported by a 3 + 1 configuration operating on a single hop, but the improvement factor would tend to be less for a multi-hop section, or a larger switch matrix[10].

## Conclusions

Many digital radio-relay systems are now in operational service on a world-wide basis, as a result of the success of the first field trial systems in the 1970s. The use of this relatively new media will grow significantly over the next decade, both in developed and developing countries, in line with the policy of most Administrations to evolve to an all-digital network.

It is now evident that digital radio systems can be expected to give a performance consistent with some of the highest quality requirements given by the CCITT for the ISDN. Moreover, the performance results coming from the many different field assessments, particularly those in trunk radio systems, indicate that these standards can now be readily achieved on most established microwave networks by using the latest system refinements.

Nevertheless, much more work is required in order to characterise the error performance of radio systems in a form usable for network management purposes, and this could be an area for the CCIR to undertake future work.

One of the most important impairments for medium- and low-capacity digital radio systems, used in regional and local network applications, is non-availability, and some indication has been given in this article of the standards that will be achieved with current systems. At present, however, there is insufficient guidance on the availability standards that are required in these parts of the network, and this must therefore be regarded as another area for future work.

Although very little representative data has so far been collected on the performance of operational low- and medium-capacity digital radio systems, it is likely that these will also be of a high grade, well within the circuit quality standards set by the CCITT.

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## Biography


Greg Hart received his B.Tech. degree in Electrical Engineering from the University of Bradford in 1971. Since graduating, he has worked in BT's Transmission System Development Department, initially on cable television system engineering. For the past five years, he has been Head of Group on digital microwave system and antenna development. He is, at present, chairman of the CEPT group dealing with the harmonisation of digital radio systems.

# Institution of British Telecommunications Engineers

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5383 *Interfacing to Microprocessors and Microcomputers.* O. Bishop (1982).

This book describes a series of practical projects, for the home, where a microcomputer system is linked to the world around it; for example, a light sensor and sound effects generator.

5384 *Sterling Engines.* G. T. Reader, and C. Hooper (1983).

This book, which gives a broad introduction to the subject of Sterling engines, is appropriate for engineers researching and developing industrial power plants, and for university courses in thermodynamics, mechanics and energy conversion.

5385 *National Giro—Modern Money Transfer.* G. Davies (1973).

This book describes the forces that led to the setting up of National Girobank, the first public-sector bank to be set up for over 100 years.

5386 *Lewis Medical, Scientific and Technical Lending Library Supplement, 1979–1981.*

5387 *Telecommunications Switching Principles.* M. T. Hills (1979).

This book is of value to engineers trained in electronic engineering, computer systems or conventional switching who need to understand the design and operation of switching systems.

5388 *Radio and Television Servicing 1981–1982 Models. Editor:* R. N. Wainwright.

5389 *Site Investigation Practice.* M. D. Joyce (1982).

This book describes, in a practical manner, all the commonly used methods of site investigation. Trial pitting, probing, shell and auger boring, and rotary drilling are described in some detail, together with *in situ* testing and specialist sampling.

5390 *Foundation Engineering—For Difficult Subsoil Conditions.* L. Zeevaert (1983).

This book gives the latest techniques for making foundations safe and preventing uneven settlement, and help on how land previously considered unsuitable for building can be developed.

5391 *Talking Computers and Telecommunications.* J. A. Kuecken (1983).

This book extensively covers computerised speech, and reviews speech digitalisation techniques, which convert sound to a digital wavetrain. New areas investigated include hardware and software techniques of the encoding form of speech synthesis and linear predictive coding.

5392 *Electronics with Digital and Analog Integrated Circuits.* R. J. Higgins (1983).

This book introduces both digital and analogue integrated circuit (IC) instrumentation. The book includes many examples of microcomputer interfacing and demonstrates applications for ICs in measurement, control, signal generation and signal processing.

5393 *Designing for Strength.* P. Polak (1982).

This book covers strength calculations for general design purposes from an introductory to an advanced level. The author explains the main stress forms first, and goes on to critically examine material tests and properties. Worked examples are included throughout the text.

5394 *Treatise on Solar Energy—Vol.1: Fundamentals of Solar Energy.* H. P. Garg (1982).

This book is the first part of a three-volume work on solar energy, which reflects its growing recognition as a viable supplementary source of energy.

5395 *Solid-State High-Frequency Power.* I. M. Gottlieb (1982).

This book is full of numerous practical examples of the new solid-state high-frequency electronics for the engineer or hobbyist.

5396 *Engineering Acoustics and Noise Control.* C. J. Hemond Jr. (1983).

The author uses real problems encountered in daily life to simplify this complex subject; he has minimised mathematical treatment to stress the art, the fundamental concepts and the continuity of acoustics.

5397 *Programming with Graphics.* G. Marshall (1983).

This book provides an up-to-date treatment of the subject, and covers the three major methods of graphic production: block, pixel and line graphics. This book is machine independent and is of great interest to all personal-computer users who use or wish to use graphics, whether in the home or in business.

5398 *Introductory Digital Electronics.* N. W. Heap, and G. S. Martin (1982).

This book is an adaptation of the Open University course, *Introductory Electronics*, and the topics covered include combinational logic, sequential logic analogue-to-digital conversions and microprocessor memory configurations.

P. W. Salter
*Librarian*

## MEMBERS ABOUT TO RETIRE

Members about to retire are reminded that they may secure life membership of the Institution at a once-and-for-all cost of £10·00. Retired members may enjoy all the facilities provided by the Institution, including a free copy of *British Telecommunications Engineering* posted to their home address.

Enquiries should be directed to the appropriate Local-Centre Secretary; members living in, or moving to, an area served by a different Centre from that to which they currently belong may find it more convenient to arrange a transfer to the new Centre's membership list before retirement in order to ensure advice on local activities.

J. Bateman
*Secretary*

**Local-Centre Secretaries**

The following is a list of Local-Centre Secretaries, to whom enquiries about membership of the Institution should be addressed. It would be helpful if members would notify any change in their own address to the appropriate secretary.

| *Centre* | *Local Secretary* | *Address and Telephone Number* |
|---|---|---|
| Birmingham | Mr. P. J. Thompson | British Telecom Midland HQ/PLT4.1, 245 Broad Street Birmingham B1 2HQ 021–262 4287 |
| Eastern (Bletchley) | Mr. D. R. Norman | British Telecom GMO/ED9.1, Telephone House, 25–27 St. John's Street, Bedford MK42 0BA (0234) 48299 |
| Eastern (Colchester) | Mr. P. M. Cholerton | British Telecom East HQ/PLG2.1.4, St. Peter's House, St. Peter's Street, Colchester CO1 1ET (0206) 89547 |
| East Midlands | Mr. D. W. Sharman | British Telecom GMO/ES3.3, 200 Charles Street, Leicester LE1 1BB (0533) 534409 |
| London | Mr. L. J. Hobson | British Telecom LCS/CM/2.1, Procter House, 100–110 High Holborn, London WC1V 6LD 01–432 2119 |
| Martlesham | Mr. G. Popple | British Telecom Research Laboratories, R5.5.1, Martlesham Heath, Ipswich IP5 7RE (0473) 644884 |
| North Eastern | Mr. R. S. Kirby | British Telecom North East HQ/PM1.3.4, Netel House, 6 Grace Street, Leeds LS1 1EA (0532) 466366 |
| Northern | Mr. L. G. P. Farmer | British Telecom GMO/I1.4, Swan House, Pilgrim Street, Newcastle-upon-Tyne NE1 1BA (0632) 613657 |
| Norther Ireland | Mr W. H. Tolerton | British Telecom GMO/EC1, Dial House, 3 Upper Queen Street, Belfast BT1 6LS (0232) 43444 |
| North Western (Manchester and Liverpool) | Mr. C. Jackson | British Telecom North West HQ/S4.4.2, Telecommunications House, 91 London Road, Manchester M60 1HQ 061–863 7777 |
| North Western (Preston) | Mr. A. Cope | British Telecom GMO/LE4, Telephone House, 170–175 Moor Lane, Preston PR1 1BA (0772) 58280 |
| Scotland East | Mr. T. L. McMillan | British Telecom Scotland HQ/T4.3.2/3, Canning House, 19 Canning Street, Edinburgh EH3 8TH 031–229 3296 |
| Scotland West | Mr. G. A. Dobbie | British Telecom GMO/EX17, Marland House, 40 George Street, Glasgow G1 1BA 041–220 2365 |
| South Eastern | Mr. S. J. Hurdle | British Telecom South East HQ/PL/EQ4.5, 52 Churchill Square, Brighton BN1 2ER (0237) 2301188 |
| South Western | Mr. D. P. Cosh | British Telecom South West HQ/Sv2.3.1, Mercury House, Bond Street, Bristol BS1 3TD (0272) 295578 |
| Stone/Stoke | Mr. J. Coulson | British Telecom GMO/LS1, Trinity Street, Stoke-on-Trent ST1 5ND (0782) 288340 |
| Wales and the Marches | Mr. D. A. Randles | British Telecom Wales and the Marches HQ/PW3.1.2.2, 25 Pendwyallt Road, Cardiff CF4 7YR (0222) 391370 |

# Forthcoming Conferences

Further details can be obtained from the conferences department of the organising body.

*Institution of Electrical Engineers,* Savoy Place, London WC2R 0BL. Telephone: 01–240 1871

**Computer-Aided Engineering**
10–12 December 1984
University of Warwick

**The ISDN and its Impact on Information Technology**
14–16 January 1985
Institution of Electrical Engineers

**HF Communication Systems and Techniques**
26–27 February 1985
Institution of Electrical Engineers

**Telecommunication Transmission**
18–21 March 1985
Institution of Electrical Engineers

**Advances in Command Control and Communication Systems: Theory and Applications**
16–18 April 1985
Bournemouth

**Antennas and Propagation (ICAP 85)**
16–19 April 1985
University of Warwick

**Installation Engineering**
5–6 June 1985
Institution of Electrical Engineers

**Software Engineering**
28–30 August 1985
Imperial College, London
*Call for papers:* Submission deadline is 15 January 1985

**Photoelectric Imaging**
10–11 September 1985
Institution of Electrical Engineers
*Call for papers:* Synopses by 5 November 1984

*Institution of Electronic and Radio Engineers,* 99 Gower Street, London, WC1E 6AZ
Telephone: 01–388 3071

**Custom VLSI for Control and Instrumentation**
6–7 November 1984
Cavendish Conference Centre, London

**Colour in Information Technology**
25–29 March 1985
University of Surrey

**Digital Processing of Signals in Communications**
22–25 April 1985
University of Loughborough

# Profiles of Senior Staff

## CHIEF EXECUTIVE TECHNOLOGY

W. G. T. JONES, B.SC., C.ENG., F.I.E.E.

William Jones was educated at Portsmouth Grammar School, and joined the telecommunications side of the Post Office in 1960 as an open-competition Assistant Executive Engineer. He then completed a four-year degree course in Electrical Engineering at Portsmouth Polytechnic.

He was one of the first telecommunications engineers to be trained in the design of integrated circuits and was closely involved in the development of the world's first digital exchange that opened in West London in 1968. He is an acknowledged authority in the telecommunications industry on switching and transmission systems, and was closely concerned with the evolution of System X.

In the late 1960s, he helped to pioneer the development of microprocessor technology and was joint author of the first public description of the use of microprocessors in telecommunications. In 1975, he began work on the introduction of digital transmission techniques in the local network; these have evolved into the integrated services digital network now being set up.

He was appointed Director of System Evolution and Standards in 1983, in charge of all aspects of research and development on networks, switching and software. He has written a number of published papers on electronic switching and holds several patents in this field.

He has been appointed British Telecom's Chief Executive, Technology, to continue effective research and development.

He is married, has one daughter, and lives in London.

## DIRECTOR: SYSTEM EVOLUTION AND STANDARDS DEPARTMENT

R. W. BRANDER B.SC., C.ENG., F.I.E.E., F.INST.P.

Robert Brander graduated from the University of Strathclyde, Scotland, with a first class honours degree in Applied Physics. He commenced full-time employment with the GEC Hirst Research Centre at Wembley, Middlesex, in 1957 and, after holding various posts, became head of its Materials Department in 1969. During this period he worked on the physics and chemistry of advanced semiconductor materials and devices; he was also responsible for directing projects on the electronic properties of high-band-gap materials, the fabrication and potential application of electroluminescent devices and cold cathodes, the advancement of various semiconductor devices and integrated circuits (ICs), and solid-state display devices.

He joined the British Telecom Research Laboratories (formerly the Post Office Research Department) in 1972 as Head of Transistor Production. Thereafter, he held various posts involving work on bipolar and hybrid technologies, non-silicon devices, and many other technologies. He became Deputy Director of Research, Advanced Technology, in 1981, with responsibility for materials research and analysis, IC technology and design and a host of other activities.

He was recently appointed Director of the System Evolution and Standards Department; he is responsible for research and development activities covering a wide range of telecommunications systems relating to the design, operation and management of the rapidly expanding telecommunications networks, and for the standards and protocols necessary to ensure the interfacing of the ever-increasing range of equipment and organisations involved in the telecommunications industry. One of the most important aspects of this work is the exploitation of recent advances in software, micro-electronics and optical technologies.

He is Editor of the international scientific journal *Solid State Electronics,* and Chairman of the Graduateship Examination Board of the Institute of Physics. He has published numerous papers, including many presented at national and international conferences and symposia, and holds external appointments on defence, industry and university committees.

# Book Review

*Introduction to Microwave Electronics.* T. C. Edwards. Edward Arnold. vi + 76pp. 48 ills. £4·95.

This book is intended to provide the background understanding of microwave electronics for undergraduates who are following systems-orientated courses.

The title's reference to microwave electronics is perhaps misleading, for it suggests a wider treatment of the subject than is achieved, even within the limitations of the book's length. In fact, the author considers the whole range of current microwave semiconductor devices. He focuses on their semiconductor properties; indicates how they can be operated as an amplifier, oscillator or modulator; and summaries their performance (often relative to alternative options) and applications. He similarly treats travelling-wave tubes, klystrons and magnetrons, among others. Unfortunately, he does not consider the transmission techniques needed to couple together such devices; and this is somewhat disappointing, given that he has previously written authoritively on this subject.

Nevertheless, technically the book is up-to-date, even though the technology is rapidly evolving, and easily readable. Indeed, the bold headings and clear diagrams enable the reader to readily scan the text for the required topic and, because of the succinct wording, rapidly assimulate the pertinent facts. Pure theory has been minimised and mathematics virtually banished without detriment to understanding.

The book should provide a handy tutorial refuge for all students of electronic engineering not in the mainstream of microwave electronics, and a source of reference for practising engineers who require briefing or simply reminding.

R. S. SWAIN

# British Telecom Press Notices

## PRESTEL FOR TELECOM AUSTRALIA

Telecom Australia has chosen Prestel, British Telecom's (BT's) viewdata system, as the basis of the new public service, *Viatel*, that it plans to start in February 1985. The contract, for GEC 4100 series computers and Prestel software, is initially worth £2M, but will increase as the Viatel system is developed and expanded. BT faced stringent competition from other countries and companies in Britain for the Viatel contract. The fact that Prestel was chosen confirms not only the system's technological superiority over other viewdata systems, but also its commercial viability.

Australia is the tenth country to purchase national videotex systems from GEC and Prestel; this was one of the main reasons why Prestel was chosen for Viatel. In addition, the Prestel standard was proven operationally and was already in use in Australia, the system could be upgraded and Australia would be able to draw on the UK's experience.

The Viatel service will be very similar to the UK's Prestel service. It will offer keyword search, which makes finding information easier for users; an electronic mail service, and simple calculation facilities at terminals; and transaction facilities that will enable users to purchase goods, book airline tickets, and undertake home banking.

More than 1000 terminals similar to Prestel are being operated by private Australian networks, and local companies will be able to supply terminals, television-set adapters and personal computers, so that they can be linked to the Viatel service. A gateway facility is also being considered for the future. This would be more efficient and offer higher speeds and lower overheads than can be offered at present.

Telecom Australia aims to give Viatel's users simple, low-cost access to information stored not only on GEC computers, but also on as many private databases as can be encouraged to join the network.

Unlike many other countries, Australia has decided not to put the Viatel service through a public trial stage. Three GEC computers capable of handling 2000 simultaneous calls each will be installed this autumn; operation will commence in February 1985.

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## INTERNATIONAL VIDEOCONFERENCE SERVICE

Engineers of the Ford Motor Company in the UK and West Germany now meet every working day for discussions without ever leaving their respective plants. This has been made possible by Europe's first commercial videoconferencing facility provided by British Telecom International's (BTI's) Business Communications Service (BCS). The BCS was established in 1983 to undertake full responsibility for running and maintaining international telecommunications systems for major business houses using BTI's knowledge of telecommunications practice in overseas countries, and is supported, on a contractual basis, by PA Computers and Telecommunications (PACTEL).

The international videoconferencing system links one fully-equipped studio in Dunton, Essex, with another in Cologne, West Germany, by satellite. The studios can be used for one hour in the morning and another hour in the afternoon, five days per week; audio conferencing is also available for six hours during the day.

The service is one of the first to use transmission capacity for videoconferencing on the European Communications Satellite (ECS1). The two identical studios are equipped to accomodate up to seven people, and up to three people in each studio can appear on the screen at any one time. Each studio has facsimile and data links, an electronic chalkboard and an array of cameras, one of which runs on a track so that different views of vehicle materials can be transmitted. With these facilities, engineers of the two countries can discuss and examine vehicle drawings, graphics, prototype parts and jointly review all forms of illustrative material, components or cars.

Ford's executives are impressed with the new service because, apart from reducing travel between the plants in the two countries, it has made meetings more productive and decision making faster.

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## PRESTEL INTRODUCES EDUCATIONAL SERVICE FOR SCHOOLS

Prestel, British Telecom's (BT's) viewdata service, has introduced a new educational service for schools. The Department of Trade and Industry has recognised the importance of information technology to schools, and is therefore helping to finance the initial development of Prestel's new service. Schools will be able to subscribe to all of Prestel's regular features, besides the special educational microcomputing applications.

The new service, which is due to begin in January 1985, is aimed at Britain's 7500 secondary schools and the teachers and advisory centres of the Local Education Authorities; it has been developed in collaboration with the Council for Educational Technology (CET) for the UK. The Micro-electronics Education Programme funded by the Department of Education and Science, the Scottish Council for Educational Technology and other educational organisations are also contributing to various parts of the service.

Prestel's new service will enable pupils to gain experience in using a modern interactive information system. Some of the special features of the service include:

(*a*) School Link, an electronic educational microcomputing service that is being produced in collaboration with *Educational Computing* magazine, and allowing schools to receive telesoftware programs direct from Prestel down the telephone line;

(*b*) advice, support and ideas to enable schools to develop their use of information technology;

(*c*) a guide to courses at all universities, polytechnics and colleges and institutes of higher education; and

(*d*) information on career opportunities.

Other developments planned for 1985 include an enhanced careers information service, and a facility for ordering educational materials and supplies.

Schools will pay reduced fees for the new educational service, and a low-cost equipment package will enable them to buy, at a substantial discount on commercial prices, everything required to convert their microcomputers into Prestel terminals.

# British Telecom Press Notices

## TWENTY YEARS OF SATELLITE COMMUNICATIONS

This summer, British Telecom (BT) celebrated more than 20 years in satellite communications with a major operation to relay live coverage of the 1984 Olympic Games in Los Angeles to television viewers in 20 European countries. The 1984 Olympics coincided with the twentieth anniversary of INTELSAT, the world's major satellite communications organisation, of which British Telecom International (BTI) is the second largest shareholder, with a stake of about 11%.

INTELSAT was established on 20 August 1964, when the original 11 members, including the UK, became signatories to the International Telecommunications Satellite consortium. BTI's earth stations at Goonhilly Downs and Madley both work to INTELSAT satellites; the 700 hours of live coverage for the Olympics contributed to the annual total of some 3000 hours of television signals handled by these stations, mostly for television news services, major international political and sporting events. BTI's new earth station in the London docklands, the London Teleport, also uses INTELSAT capacity for transmission of cable television programmes.

There are now 109 member countries of INTELSAT; these operate about 800 aerials for international and domestic communications, all working to one or more of the 15 INTELSAT satellites now in geostationary orbit 36 000 km above the equator.

BTI routes 65% of its intercontinental telephone traffic via INTELSAT satellites. A total of 11 aerials at Goonhilly Downs and Madley connect British telephone users to more than 80 countries by satellite. Services are provided at all times of the day for television stations and news agencies world-wide.

At the start of the satellite era, Aerial 1 at Goonhilly Downs was used to conduct pioneering trials when TELSTAR, the first communications satellite, was launched in 1962. Because TELSTAR had a low orbit, it could be seen from Goonhilly Downs for only about half an hour at a time, and Aerial 1 had to be nimble enough, despite its 1100 tons, to follow it round as it circled the earth.

INTELSAT started the world's first commercial service with EARLYBIRD (INTELSAT I) in 1965; this was the first geostationary satellite. INTELSAT could carry the equivalent of 240 telephone calls or one television channel. The latest generation of INTELSAT V satellites can handle 12 000 simultaneous telephone calls, together with two television channels. A new generation of INTELSAT VI satellites will be even more powerful.

In 1965, Goonhilly Downs was the first European earth station to transmit 'live' colour television signals, via INTELSAT I, to North America. A year later, Goonhilly Downs received the first live television signals from Australia, via an INTELSAT II satellite. BTI opened a second earth station at Madley in 1978 to cope with the fast growth in international communications; the station now has five aerials. New techniques in both satellites and dish aerials have meant a new era of smaller-sized aerials. At Goonhilly Downs and Madley, the aerials are up to 32 m in diameter: at the London Teleport, they are 13 m in diameter. BTI's SatStream service for business uses even smaller dishes, from 3·7–5·5 m in diameter, located on or close to a customer's premises.

The first SatStream service, which opened (to Canada) earlier this year, operates to an INTELSAT V satellite, and is the first international service of this kind. SatStream uses digital transmission techniques to enable customers to send and receive computer data, text and facsimile at very high speeds. Another 'first' was achieved when international videoconferencing, using INTELSAT satellites, was introduced earlier this year.

Next year, BTI will be introducing time-division multiple access (TDMA), a system that will use current and planned INTELSAT satellites and allows greater volumes of telephone and non-speech traffic to be transmitted. At both Goonhilly Downs and Madley, TDMA terminals and equipment are nearing completion.

INTELSAT satellites transmitted about 5500 hours of television of the Los Angeles Olympic Games from USA earth stations to virtually all INTELSAT member countries.

## PICTUREPHONE

A versatile new wall telephone that gives users freedom to change its appearance as they please has been designed for British Telecom (BT) by the industrial design group Jones Garrard Ltd. of Leicester and Bristol.

The new telephone, called the *Picturephone*, is the result of an imaginative approach to design by BT.

The Picturephone comprises a compact telephone handset and keypad fitted into the corner of a box picture frame. The remaining area of the frame can be used to display any photograph or illustration beneath a perspex cover to create a decorative effect. It can also be used to display notices, essential telephone numbers or an advertisement. The perspex cover can be replaced with a cork pin board for messages, or a mirror board; these are provided as optional accessories.

The new telephone was designed in response to BT's brief for a new approach to design. In order to find attractive, modern designs for wall or desk telephones, a project was set up involving four selected UK design consultants, who were asked to consider how the existing range could be redesigned. For the design exercise, the brief was that no detail of the existing mechanisms available was to be touched, but a bracket or two might possibly be modified. The Picturephone was the first new idea as a result of the design exercise to go into production. It is being manufactured at BT's factory in South Wales with the latest electronic components including a memory for re-dialling.

The Picturephone

# British Telecom Press Notices

## REPAIR SERVICE CONTROL CENTRES

Earlier this year, British Telecom (BT) set up the first of its new-style repair service control centres (RSCs) at Croydon in Surrey; by the end of 1984, a further five RSCs, serving 54 telephone exchanges, will have been linked to the system.

The inauguration of the new system at Croydon marks the latest phase in the multi-million pound scheme that BT is undertaking over the next three years to make its telephone fault repair service more efficient. In future, customers dialling 151 to report faults on the telphone network will be answered increasingly by customer service officers (CSOs), staff that have been specially selected and trained for this work. These officers are located in special offices covering several telephone exchanges.

The CSOs have all the benefits of modern technology to help them do their work quickly and efficiently. Their work stations are equipped with visual display units (VDUs) connected to minicomputers and remote line testers (RLTs). When faults are reported, the CSOs can call up and consult customers' records on their VDUs, and then carry out a preliminary series of simple tests. Subsequently, the CSOs pass details of the reported faults automatically to testing officers, who use the system to carry out more complex diagnostic tests. As RLTs become widely available, technicians will seldom have to visit telephone exchanges to carry out tests. The results of the tests form part of the fault reports, which are transmitted through the system to the engineers that distribute work to field-repair and exchange-maintenance staff.

The new fault-repair scheme should enable faults in the network to be identified much more easily and quickly. In some cases, the computer can calculate the length of time that it could take to repair a particular fault; and this enables CSOs to arrange for technicians to call on customers at specific times to repair faults.

The new scheme has obvious advantages. The use of computers minimises paperwork so that experienced technicians are freed to concentrate on repairs instead of undertaking initial testing. Moreover, as the faults are reported, technicians working in the locations concerned can be directed to respond quickly to rectify the faults.

The system can also be used to make routine tests overnight of customers' lines; therefore, developing problems can be located before the customer becomes aware of their existence. Furthermore, this will render visits by technicians to customers' premises unnecessary.

Between 1983 and 1984, BT handled more than 21 million fault reports; the new scheme is designed to help BT continue the steadily improving quality of its repair service to customers. In the past five years, the number of faults cleared by the next working day has risen from 50% to over 90%; BT's target is 95%.

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## HIGH-SPEED BRITISH TELECOM LINKS FOR PAYE NETWORK

British Telecom's (BT's) investment in a national advanced digital communications network is enabling the Inland Revenue to modernise its Pay as You Earn (PAYE) operations.

Computerisation of PAYE (COP) relies on BT's national network of high-speed digital transmission services, which is being continually expanded.

This project is one of the biggest of its kind that BT has undertaken for an individual customer and one of the first commercial applications of KiloStream Plus, the installation of the KiloStream multiplexer in the customer's premises rather than in the KiloStream exchange. The project was planned by BT Wales and the Marches' Major Projects Group, which is also co-ordinating the national link-up.

By the end of 1986, nearly 800 circuits dedicated to PAYE operations will provide direct computer-to-computer and computer-to-district-office links between 600 district tax offices, 11 regional processing centres, a National Development Centre at Telford and other Inland Revenue main-frame computers around the country. The provision of operational links is already under way in a pilot scheme. Between now and the end of 1988 an average of one Inland Revenue district office will be connected to the national digital network every day. The next regional processing centre, at Peterborough, will join the network in January 1985, with others following at nine-weekly intervals.

British Telecom's digital services give the Inland Revenue:

(*a*) top-level security of data transmission;
(*b*) speed, with transmission rates of up to 48 kbit/s between processing centres;
(*c*) flexibility, since the network can be adapted rapidly to provide additional communications links and alter existing ones; and
(*d*) economy, because the service can carry speech and data at the same time.

The COP system is provided through a secure and comprehensive network of high-capacity coaxial cable, transverse-screened cable, optical-fibre cable, and microwave radio.

The design of the network enables complete flexibility to be achieved; calls can be routed through alternative channels if a circuit fails.

High-speed transmission is achieved through digital transmission techniques. It is cheaper, faster and more efficient than conventional analogue transmission.

The COP package, which will cost £14M over 13 years (the Inland Revenue's accounting period), highlights the savings available from digital transmission. An equivalent analogue network including modems would have cost more than £21M over the same period.

The KiloStream Plus network will, by connecting Inland Revenue offices across the country, give direct communications between mainframe computers and between computers and district offices as well as give other possible applications such as voice traffic, facsimile, slow-scan television and videoconferencing all condensed in a single multiplexed network, the key to keeping costs down.

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Full membership of FITCE in the UK is available only through IBTE. Members and Affiliated Members of IBTE who hold a University science degree or who are Chartered Engineers may join through the FITCE Group of IBTE. The annual subscription for 1984/85 has been fixed at £5·00; this covers local administration expenses as well as the *per capita* contribution to FITCE funds, and thus ensures that no charge proper to FITCE affairs will fall upon the general membership of IBTE. Membership forms are available from your Local-Centre Secretary (see p. 228 of the October 1983 issue of this *Journal*) or direct from the Assistant Secretary (FITCE), Mr. P. A. P. Joseph, BTHQ/TES 3.1.3.2, Room 314, Broad Street House, 55 Old Broad Street, London EC2M 1RX; Tel: 01-588 8970.

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# Notes and Comments

### INCREASE IN SUBSCRIPTION RATES

The Board of Editors regrets that the price of the *Journal* will increase from the January 1985 issue. The new price will be £1·00 (£1·50 including postage and packaging); annual subscription: £6·00 (including postage and packaging) (Canada an the USA $10·00). The special price to British Telecom and British Post Office staff remains at 51p per copy.

### APOLOGY

The Board of Editors apologises for the late publication of the July 1984 issue of the *Journal*. This was, in part, to allow for the three articles on performance requirements for digital networks (pages 92, 99 and 108), which first appeared in *The Radio and Electronic Engineer*, to be revised to take account of the final CCITT SG XVIII meeting during the 1980–84 study period.

### EDUCATIONAL PAPERS IN THE SUPPLEMENT

The *Supplement* included with this issue of the *Journal* contains the first part of a revised version of one of the series of *Educational Pamphlets* originally published by British Telecom—*Microcomputer Systems*. It is anticipated that the second part will be published in the January 1985 issue.

The Board of Editors first experimented with the publication of this type of material in the October 1982 issue, when the *Educational Pamphlet* entitled *Field-Effect Transistors* was reproduced. It was hoped that these papers would complement the articles published in the *Journal* and the question/answer material traditionally published in the *Supplement*, and, hopefully, broaden the appeal of the *Journal*. The Editors now hope to include similar educational papers on a more regular basis.

Question and answer material, aimed at BTEC and SCOTEC students will, of course, continue to be included in the *Supplement*.

### CONTRIBUTIONS TO THE JOURNAL

Contributions to *British Telecommunications Engineering* are always welcome. In particular, the Board of Editors would like to reaffirm its desire to continue to receive contributions from Regions and Areas, and from those Headquarters departments that are traditionally modest about their work.

Anyone who feels that he or she could contribute an article (short or long) of technical, managerial or general interest to engineers in British Telecom and the Post Office is invited to contact the Managing Editor at the following address: *British Telecommunications Engineering*, LCS/P5.1.1, Room 704, Lutyens House, Finsbury Circus, London EC2M 7LY.

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# British Telecommunications Engineering

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Alternatively, the panel below, or a photocopy, can be used (only by members) to borrow any book listed in the catalogue or in the updating lists published in recent issues of the *Journal*. The form should be sent to the Librarian at the address above; a self addressed label must be enclosed.

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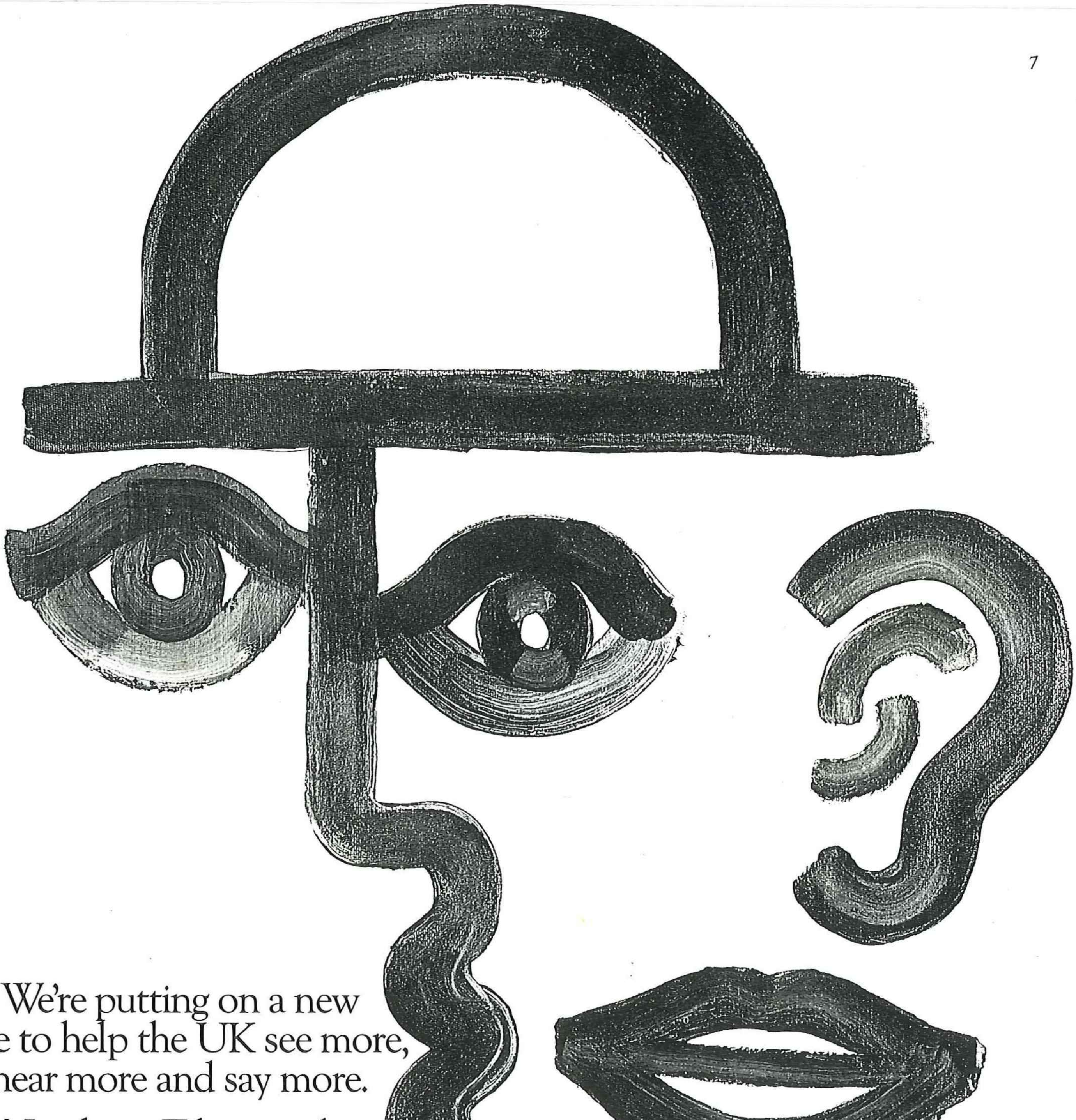

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